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(12) **United States Patent**
Suominen

(10) **Patent No.: US 6,631,256 B2**
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(54) **SIMPLIFIED HIGH FREQUENCY TUNER AND TUNING METHOD**

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(73) Assignee: **University of Washington**, Seattle, WA (US)

(*) Notice: Subject to any disclaimer, the term of this patent is extended or adjusted under 35 U.S.C. 154(b) by 0 days.

This patent is subject to a terminal disclaimer.

(21) Appl. No.: **10/032,526**

(22) Filed: **Oct. 27, 2001**

(65) **Prior Publication Data**

US 2002/0090924 A1 Jul. 11, 2002

Related U.S. Application Data

(60) Continuation of application No. 09/317,781, filed on May 24, 1999, now Pat. No. 6,427,068, which is a division of application No. 08/713,761, filed on Sep. 13, 1996, now Pat. No. 5,937,341.

(51) **Int. Cl.**[7] **H04B 1/10**
(52) **U.S. Cl.** **455/302**
(58) **Field of Search** 455/302

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(57) **ABSTRACT**

An RF tuner and tuning method employs analog quadrature mixing with a coarse-stepwise tunable local oscillator to a near-baseband passband region, followed by A/D conversion of the I and Q signals, correction of phase, group delay, and amplitude errors, image rejection, and translation to baseband by (1) fixed frequency translation, (2) stepwise channelized translation, or (3) essentially continuously variable tuning over a given digital tuning range. The near-baseband passband region is sized and located such that alternating image rejection provides non-redundant and complete tuning coverage of a desired high frequency spectrum with a local oscillator step size equal to about twice the digital tuning range or about twice the number of channels digitally stepwise tunable times the channel width, effectively doubling the typical local oscillator step size. The digital tuning is preferably performed by a continuously variable bandpass decimating filter with aliasing to within R_D of baseband, where R_D is the filter's output sampling rate, followed by fine-shifting to baseband by digital complex mixing. Demodulation is then accomplished according to signal type.

Image rejection and phase error and gain correction are preferably performed with a modified type III Hilbert transform pair with 90°±CF phase change and variable gain. The near-baseband passband is preferably centered at R/4 where R is the sampling rate entering the Hilbert transform.

17 Claims, 11 Drawing Sheets

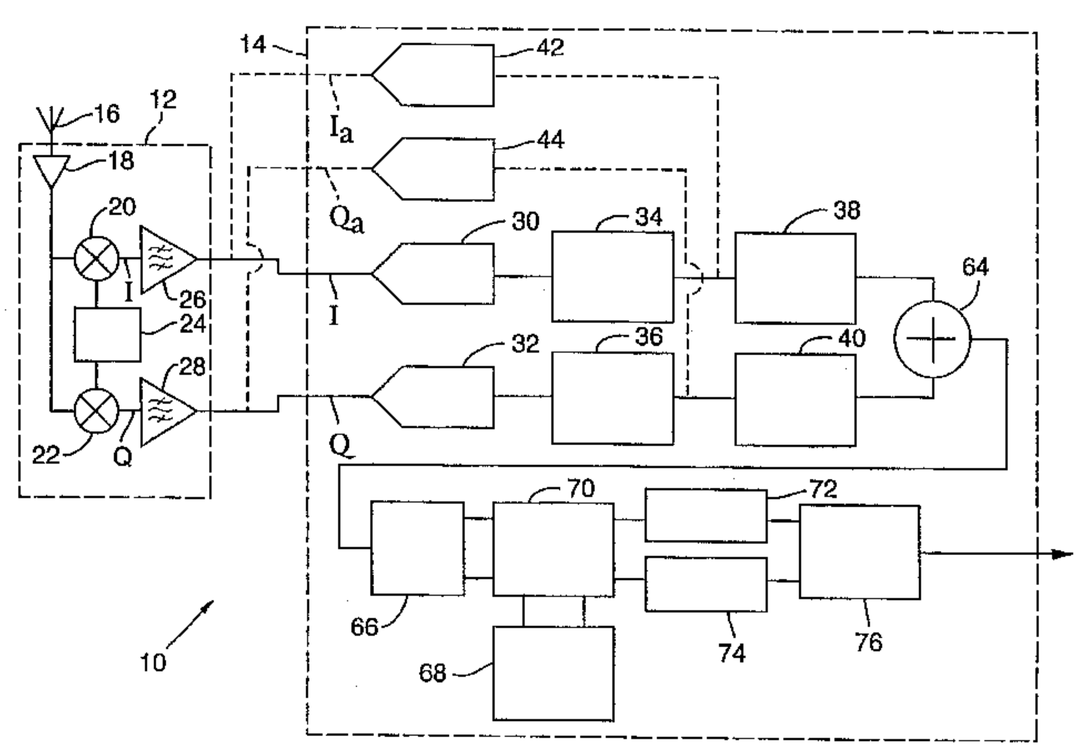

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FIG. 1

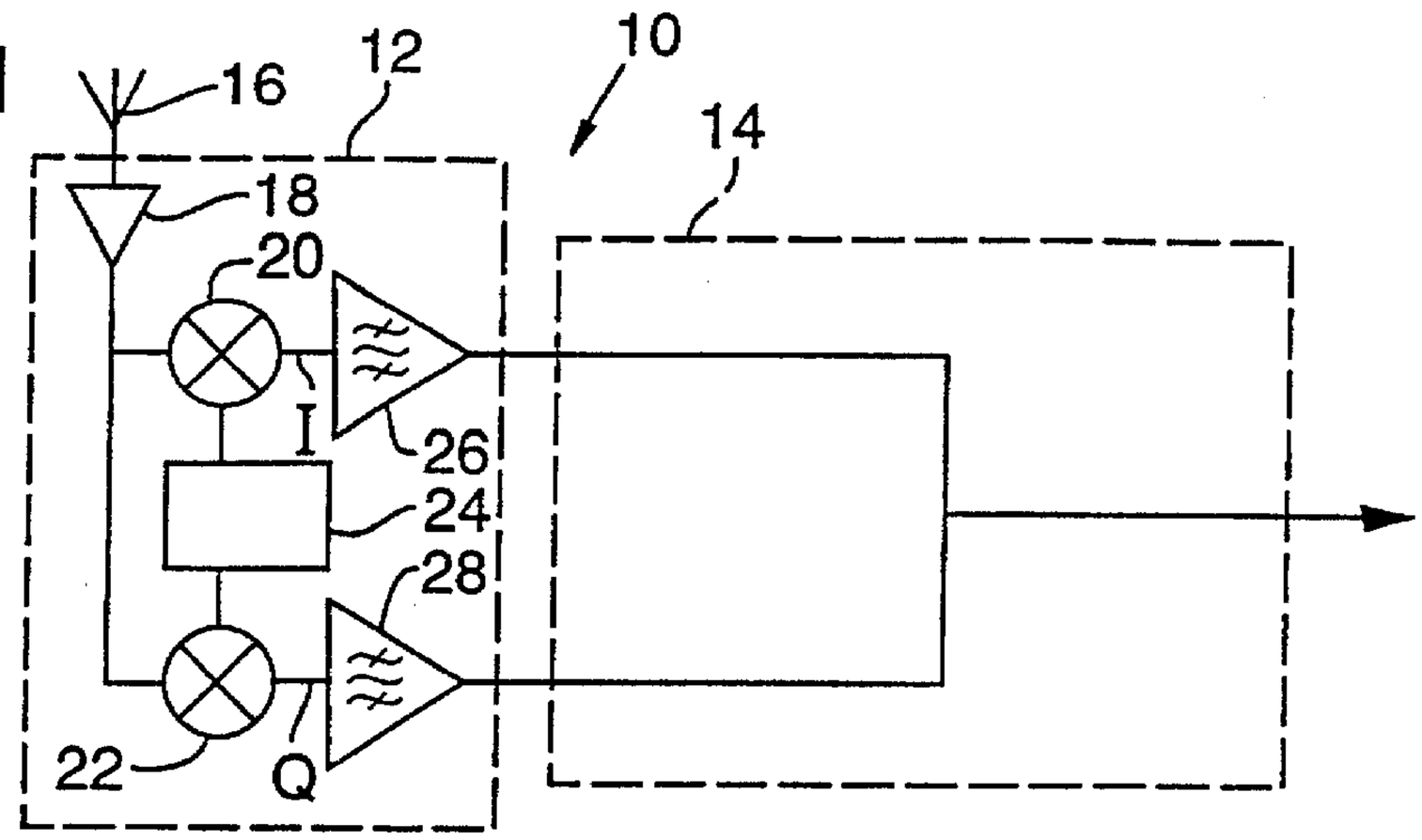
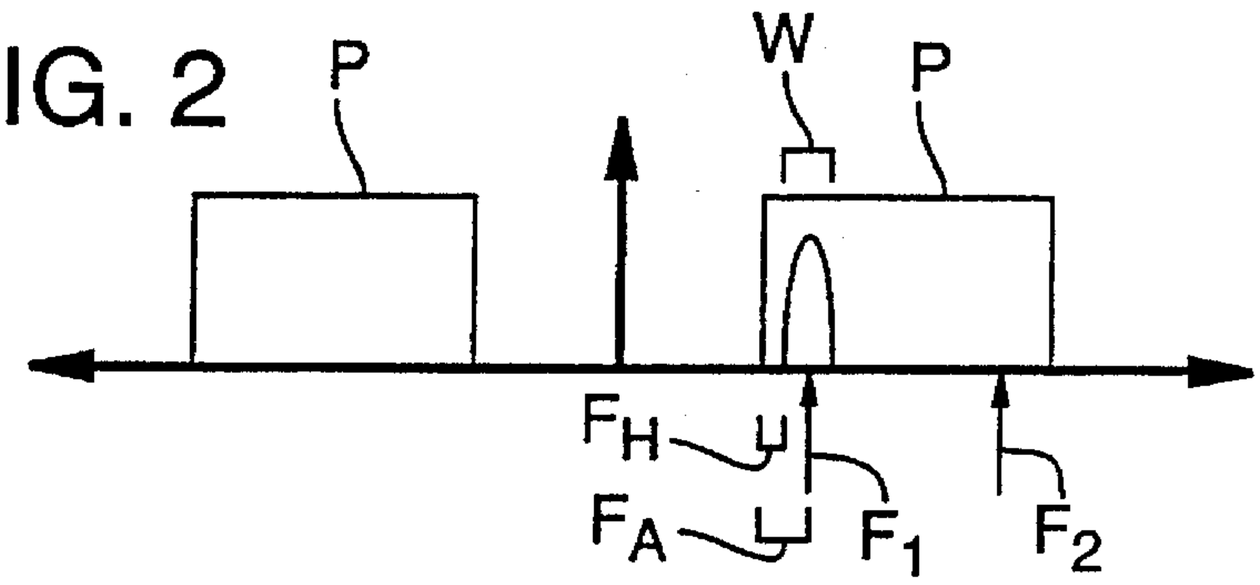


FIG. 2



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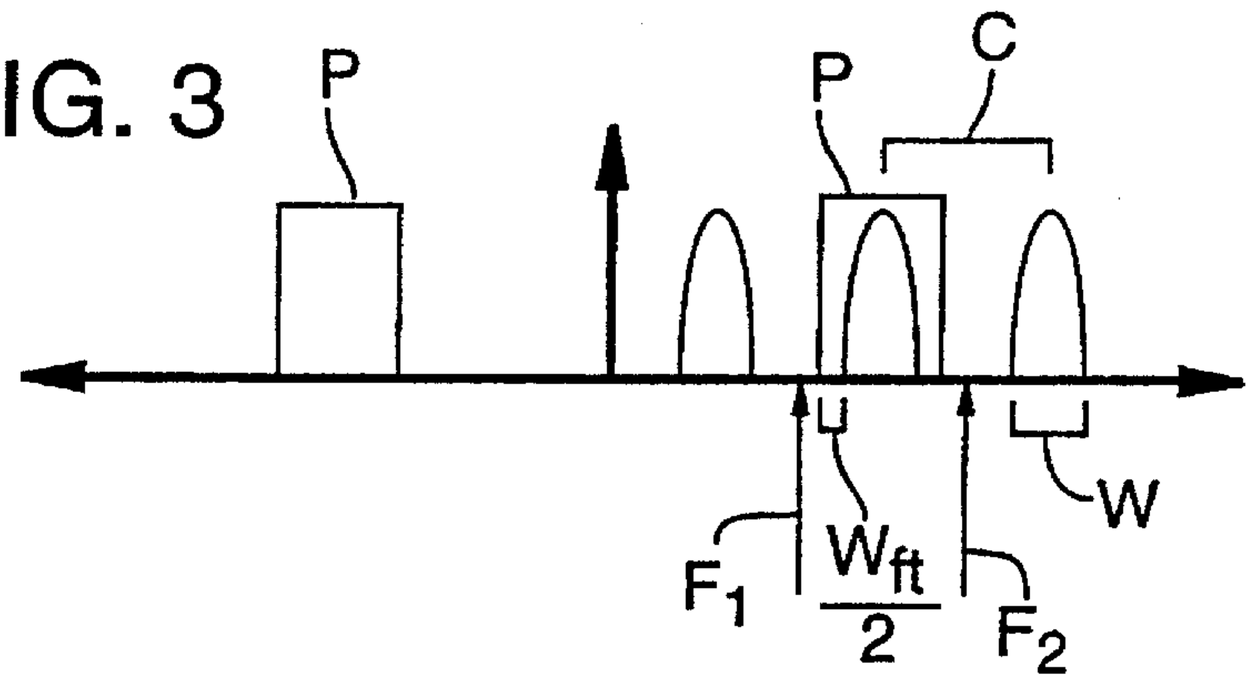

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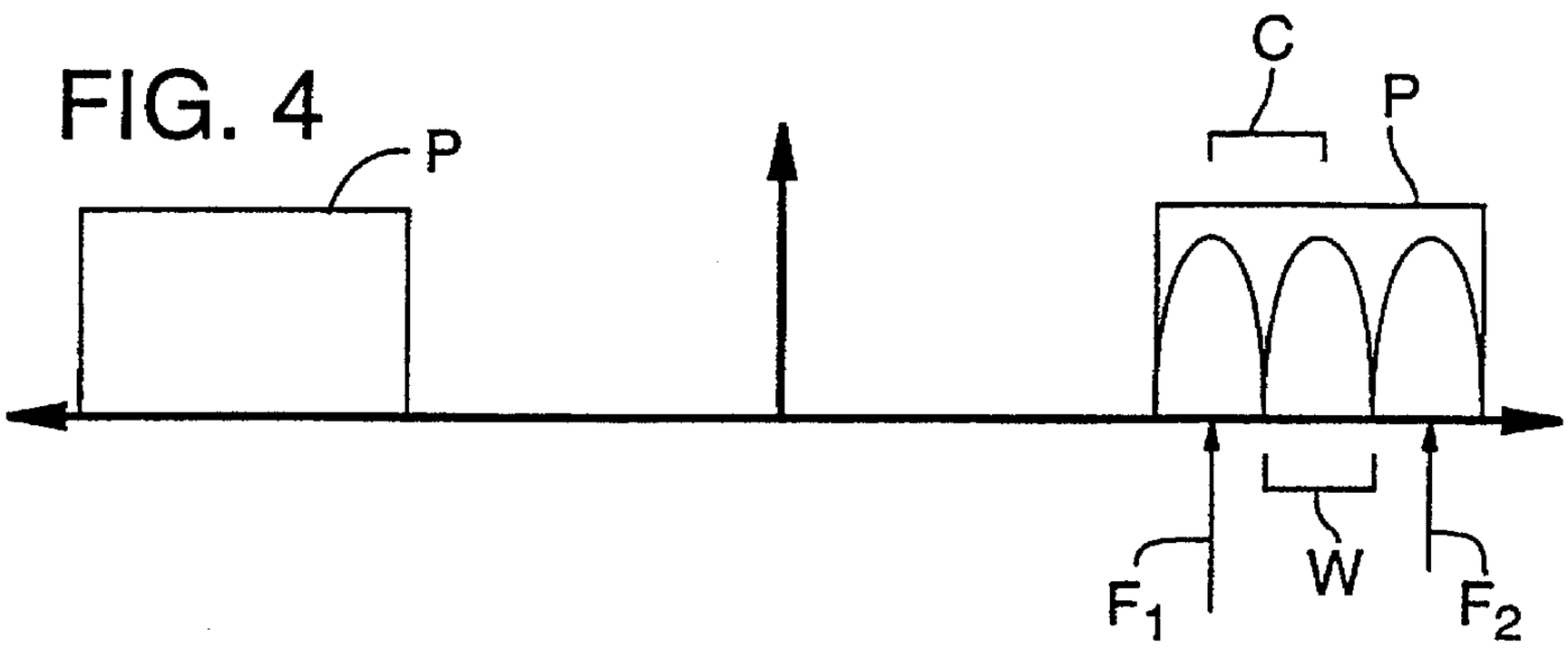

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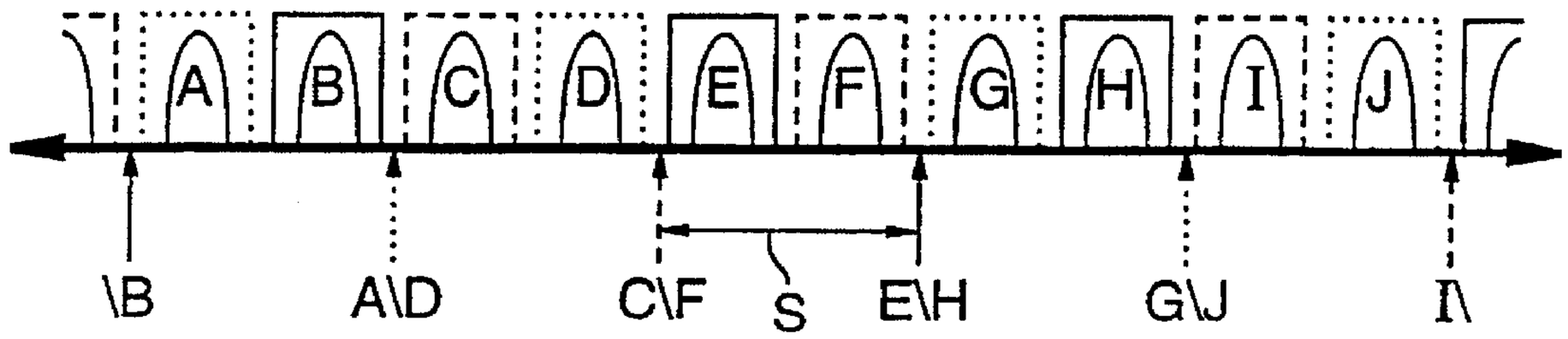

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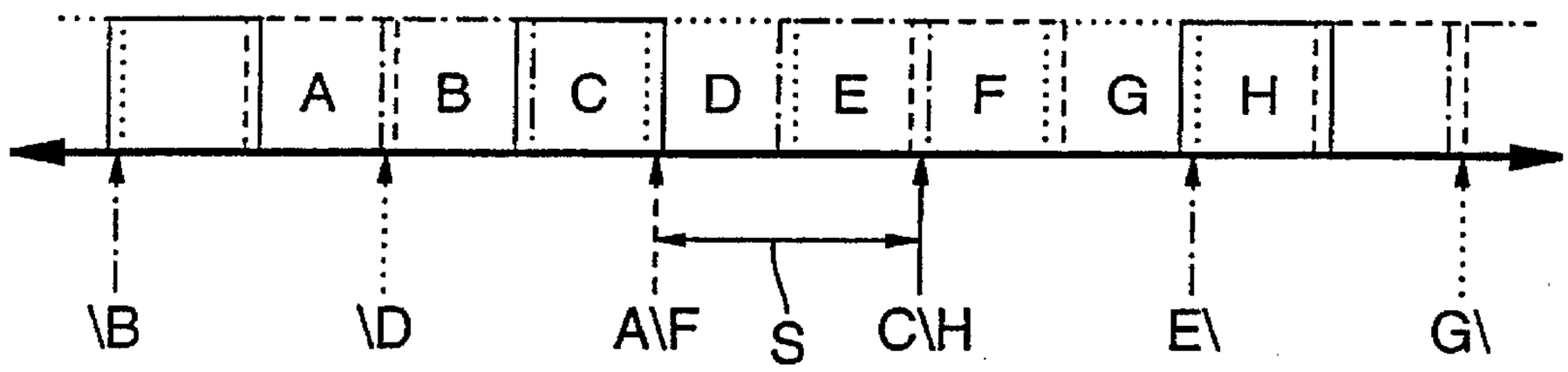

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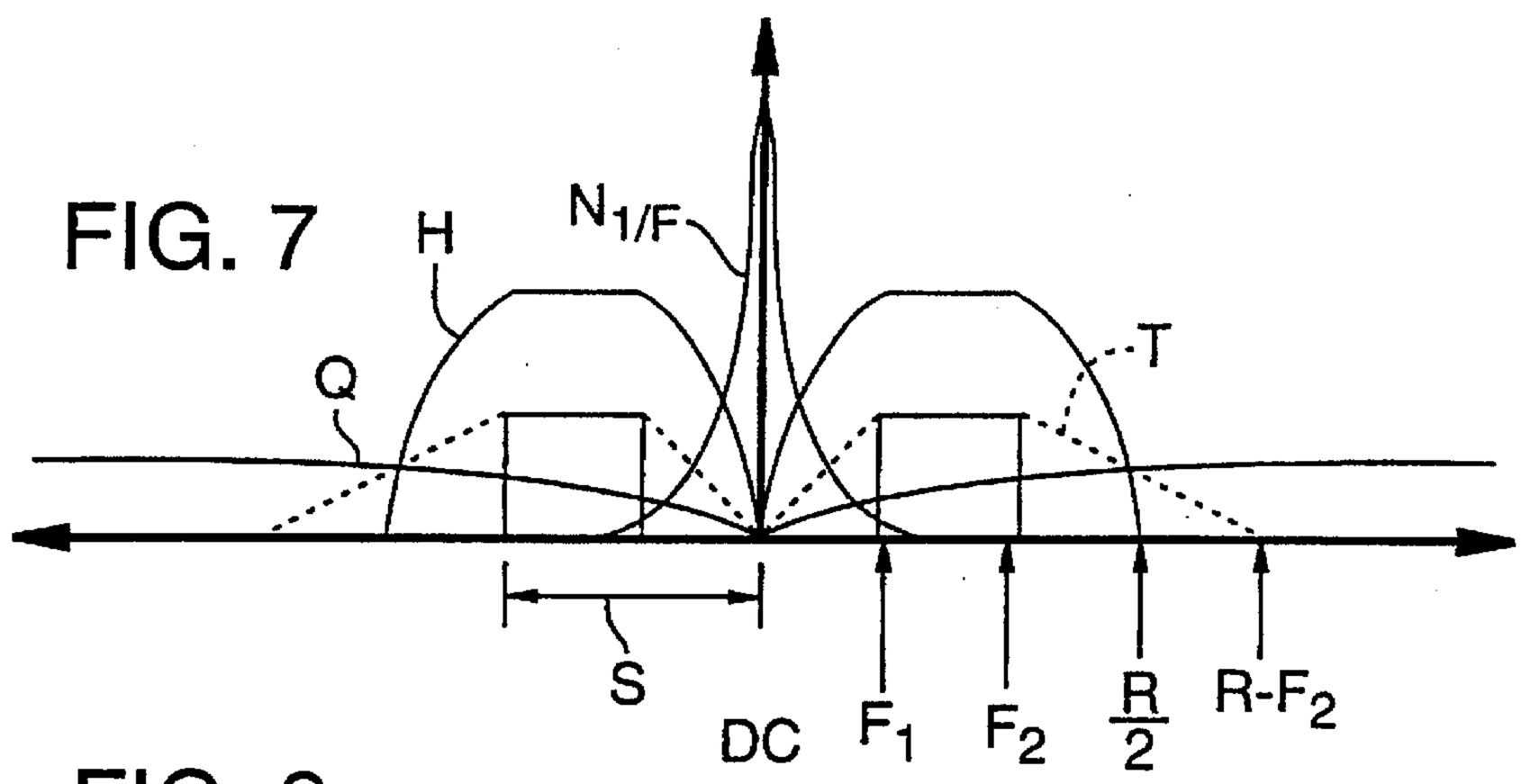

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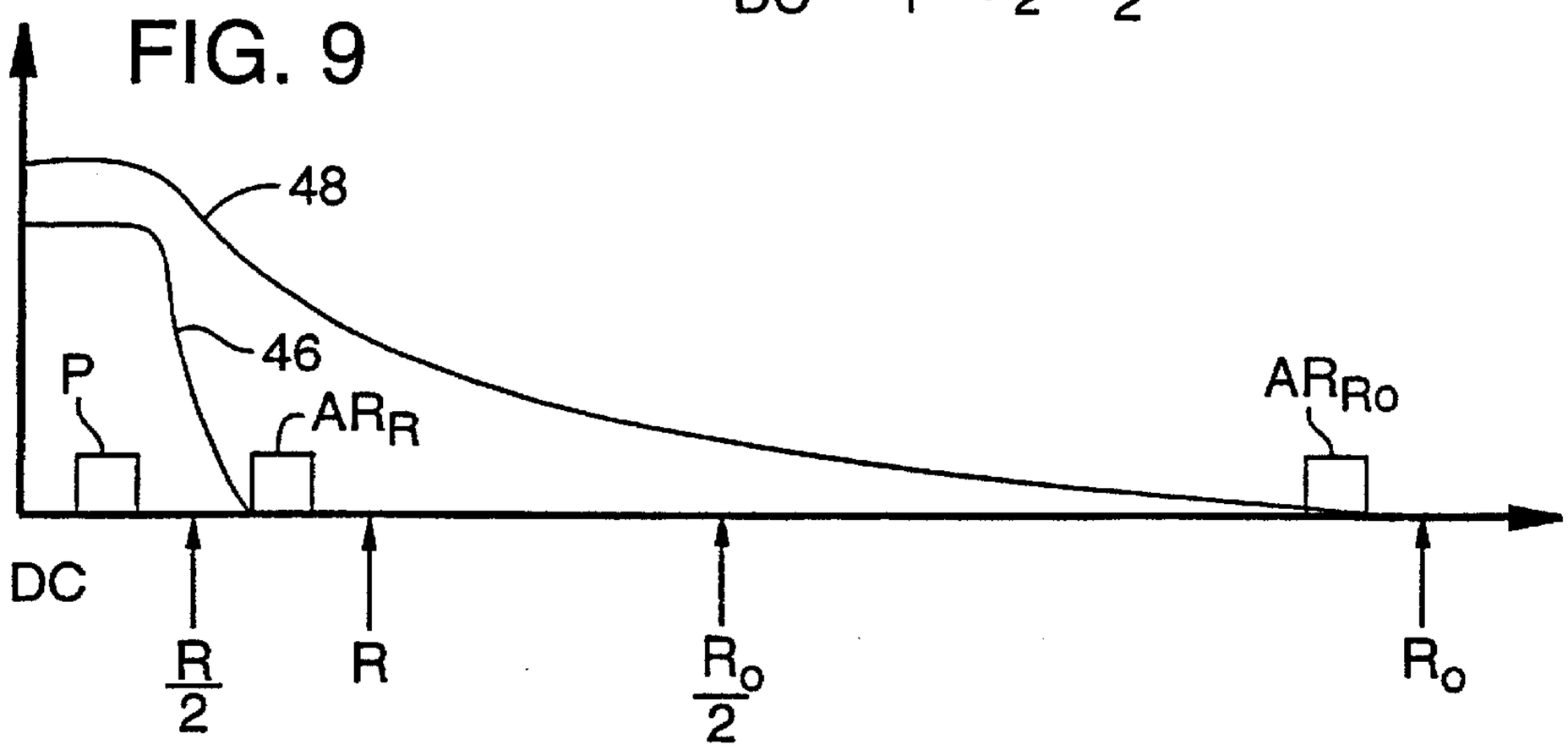

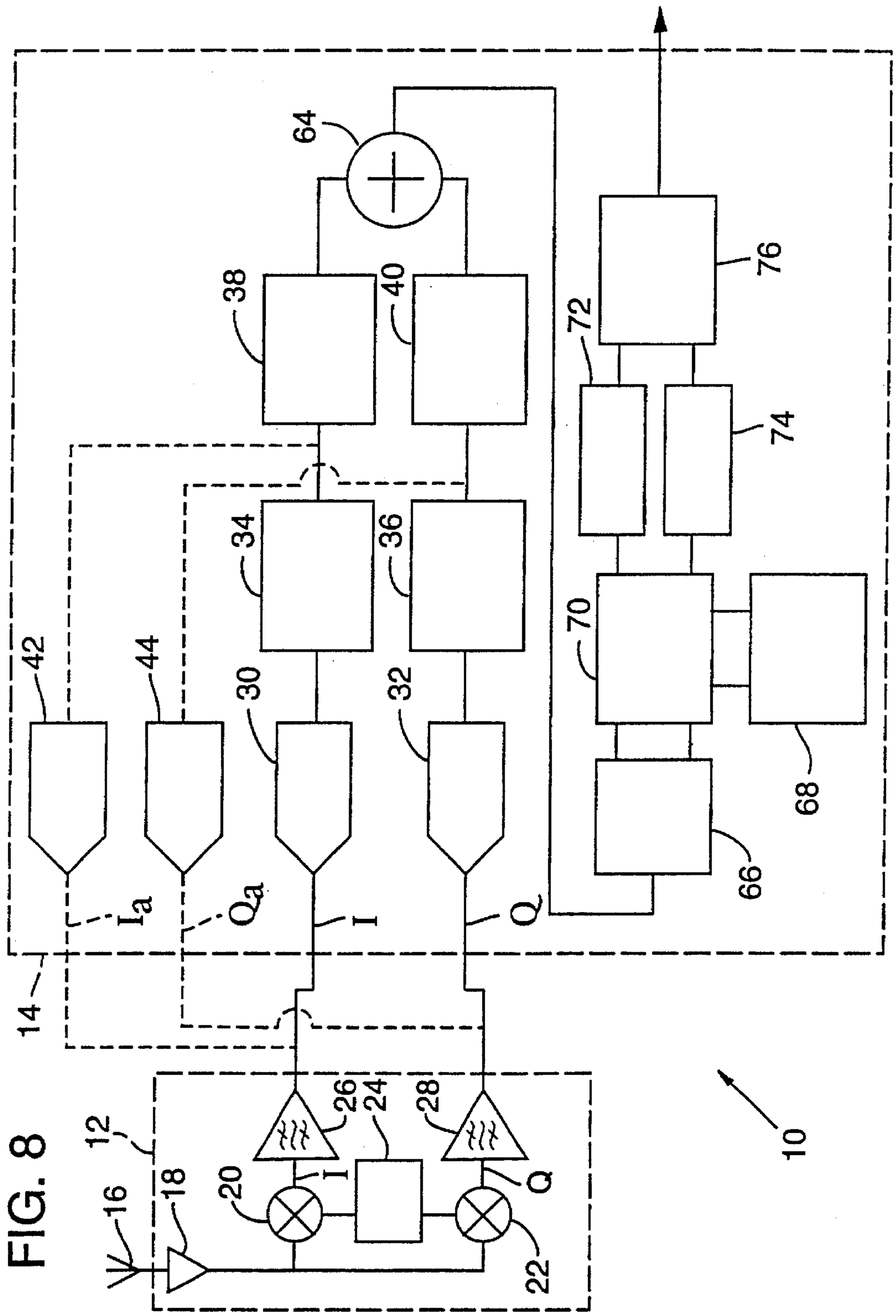
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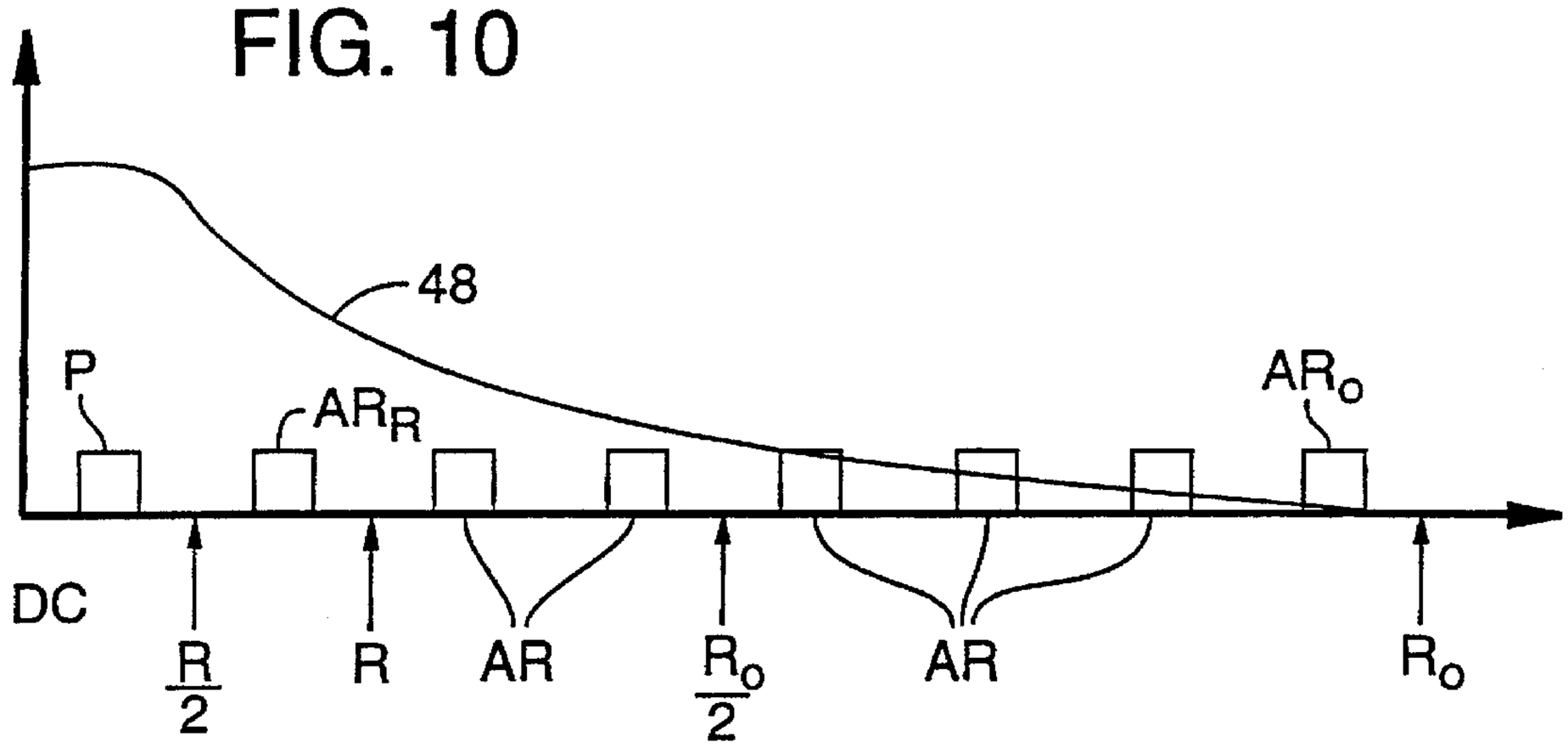
72
74
76
I
Q
Ia
Qa

FIG. 8

FIG. 10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AR_R
AR_O
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R
AR
$\frac{R_O}{2}$
AR
R_O

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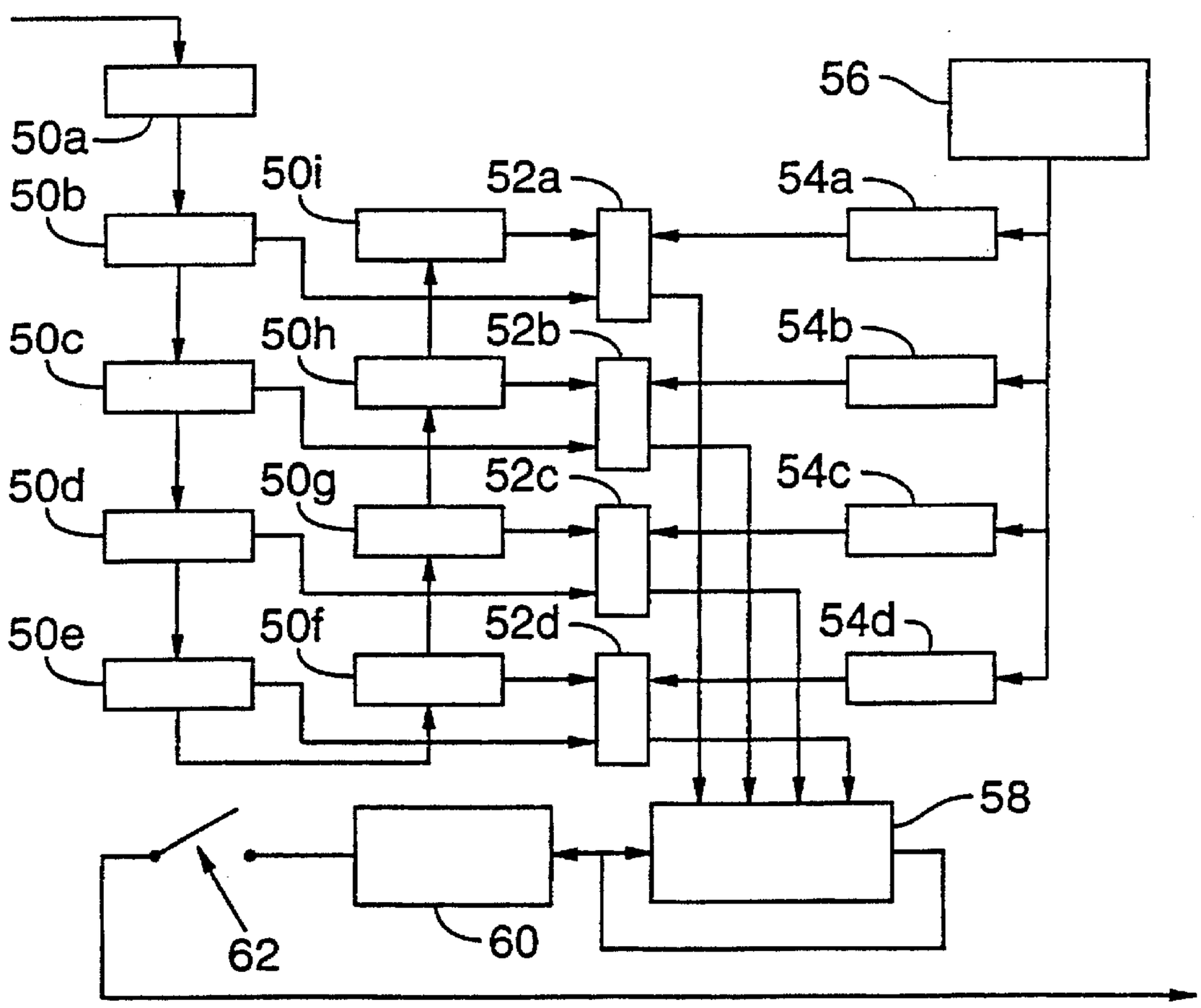
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50i
52a
54a
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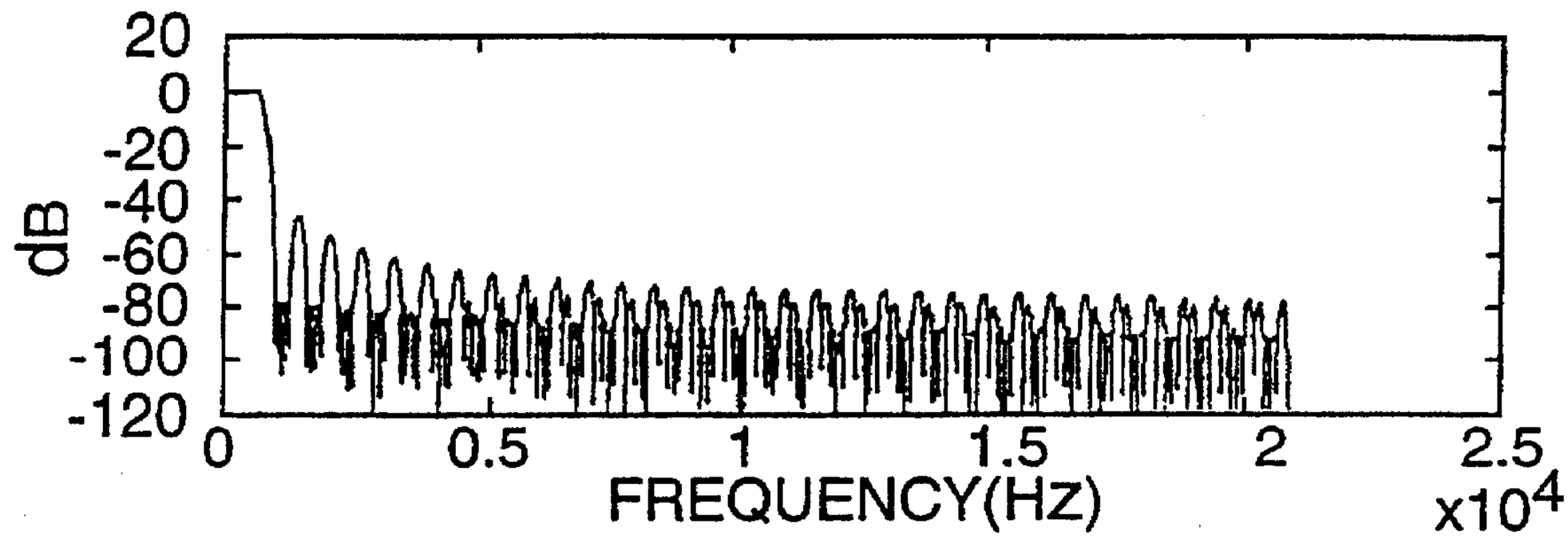
FIG. 12



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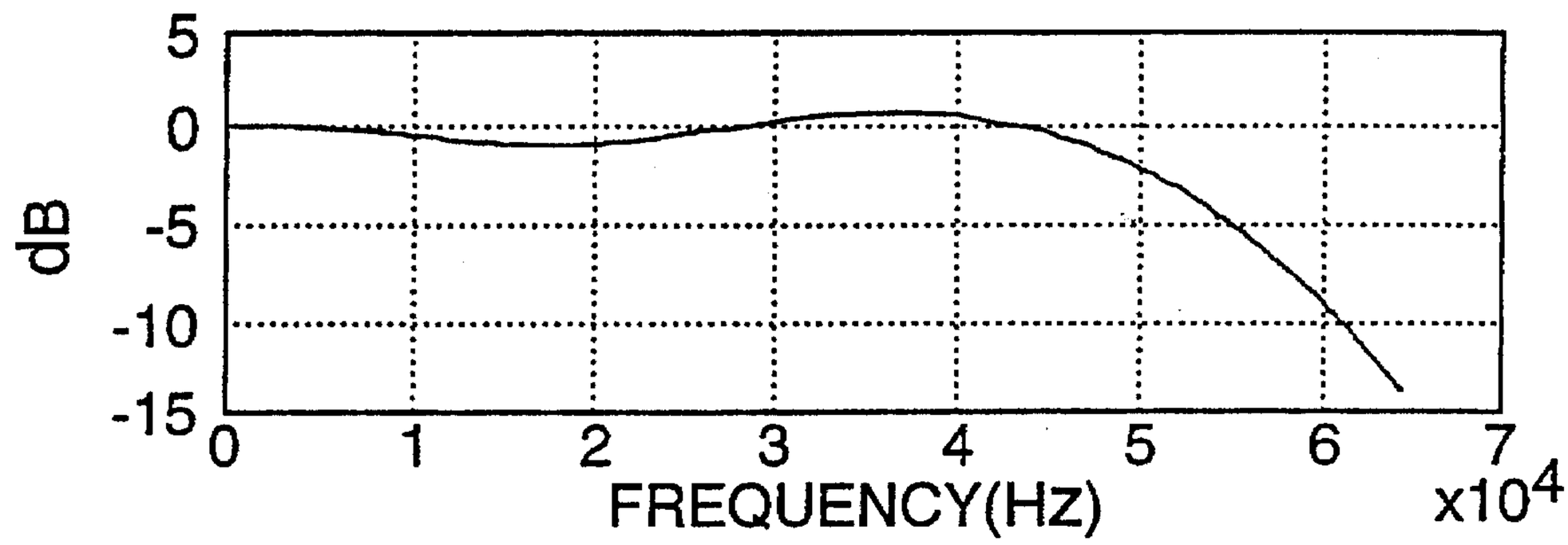

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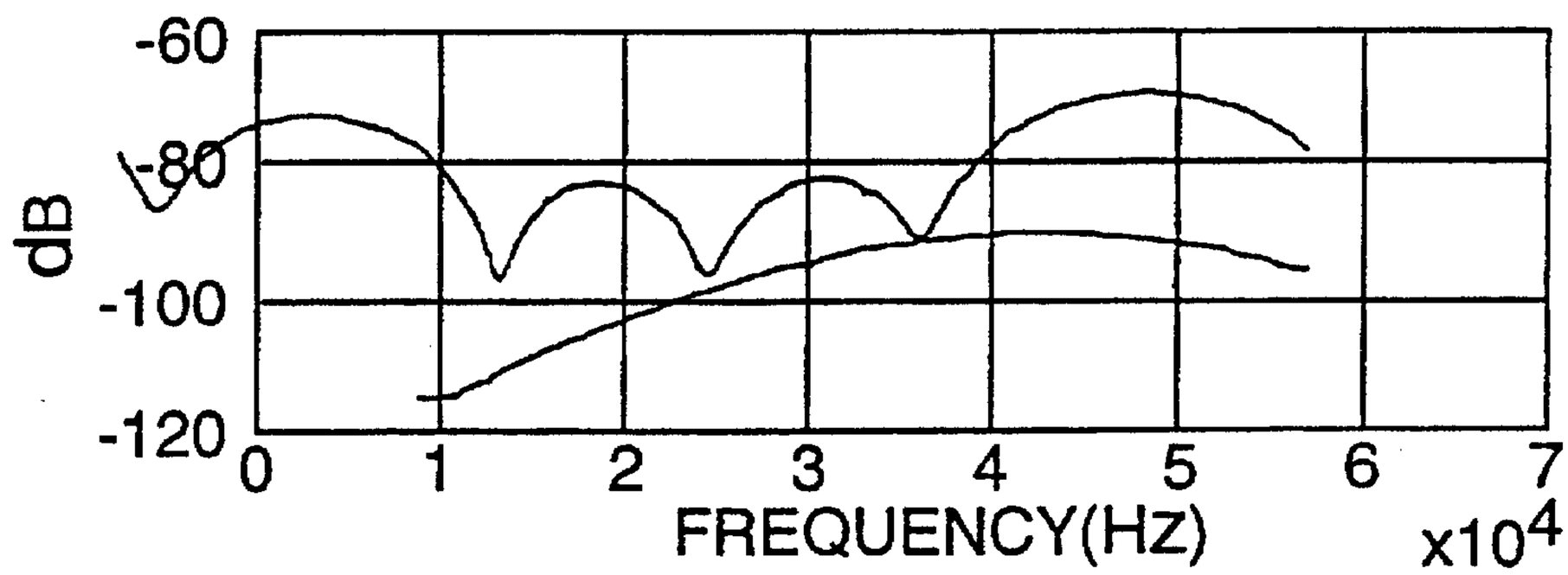

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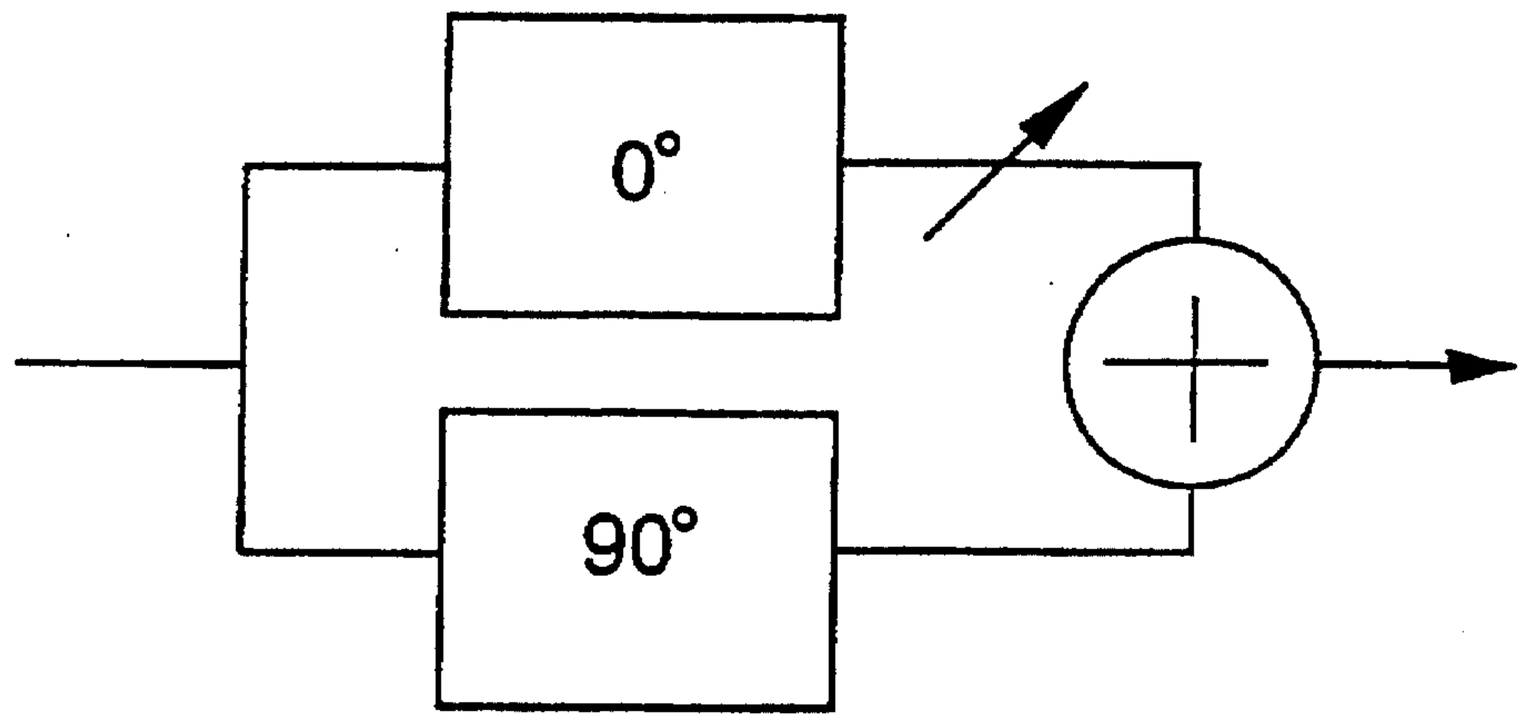

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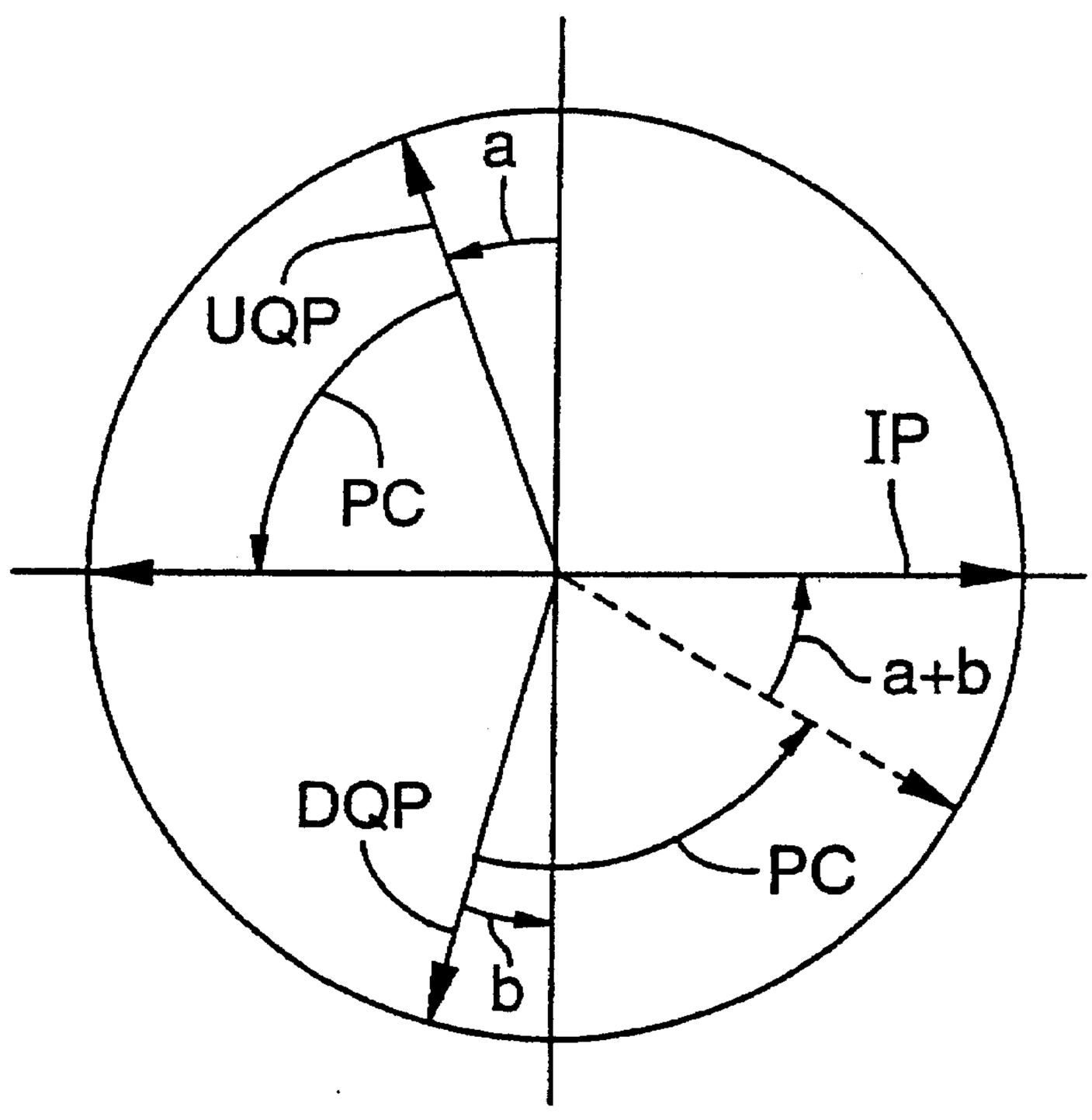

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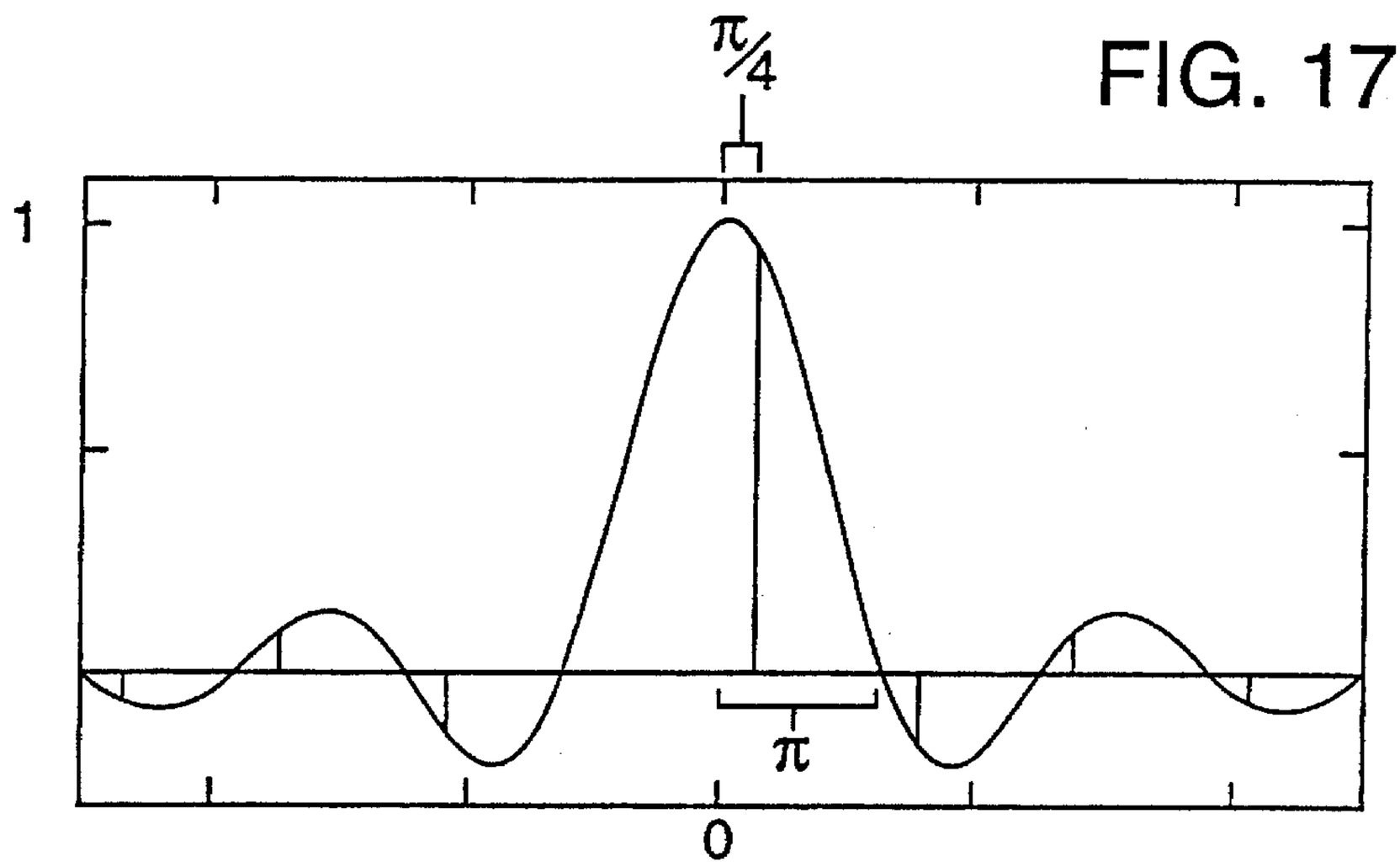

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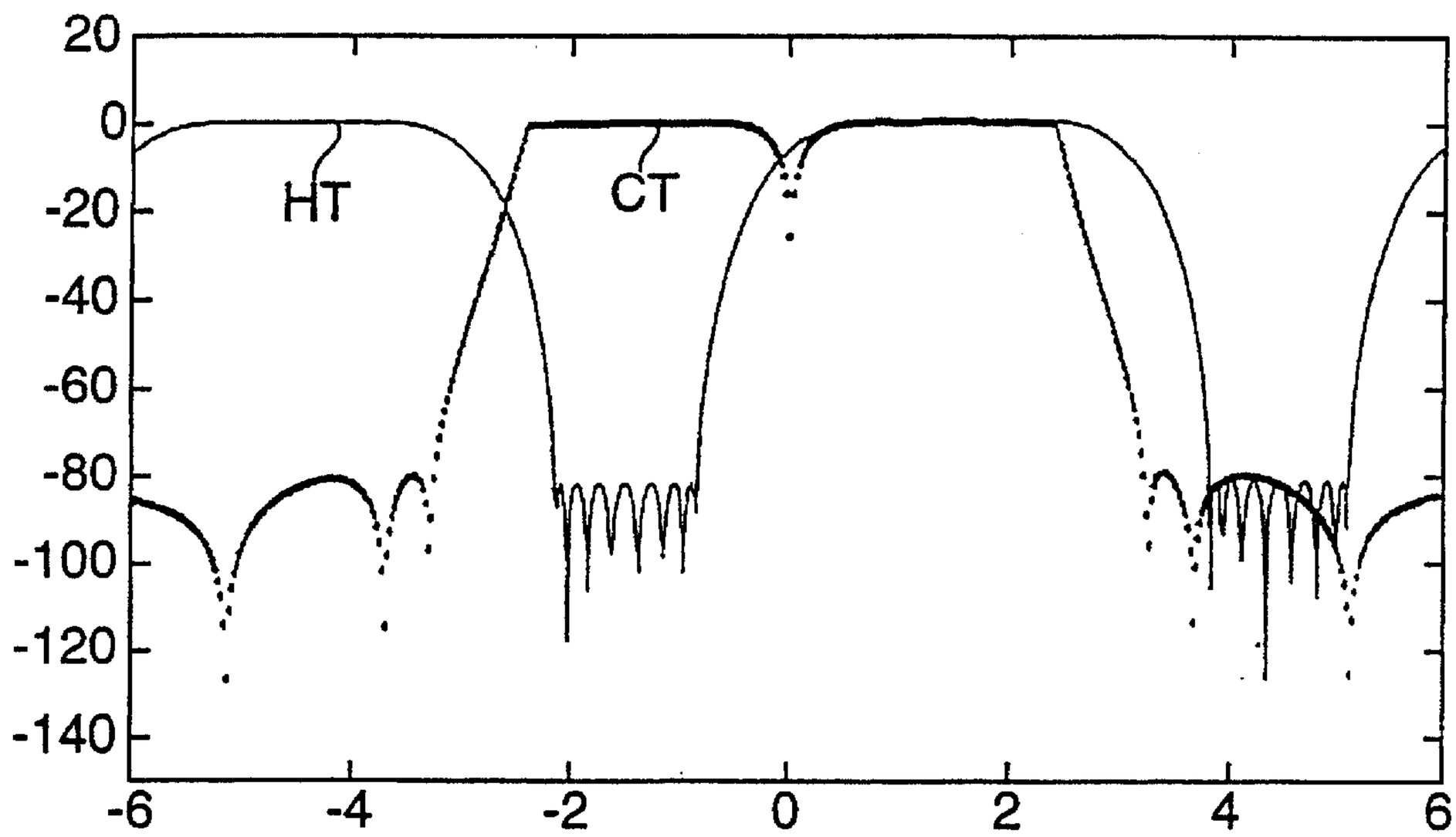

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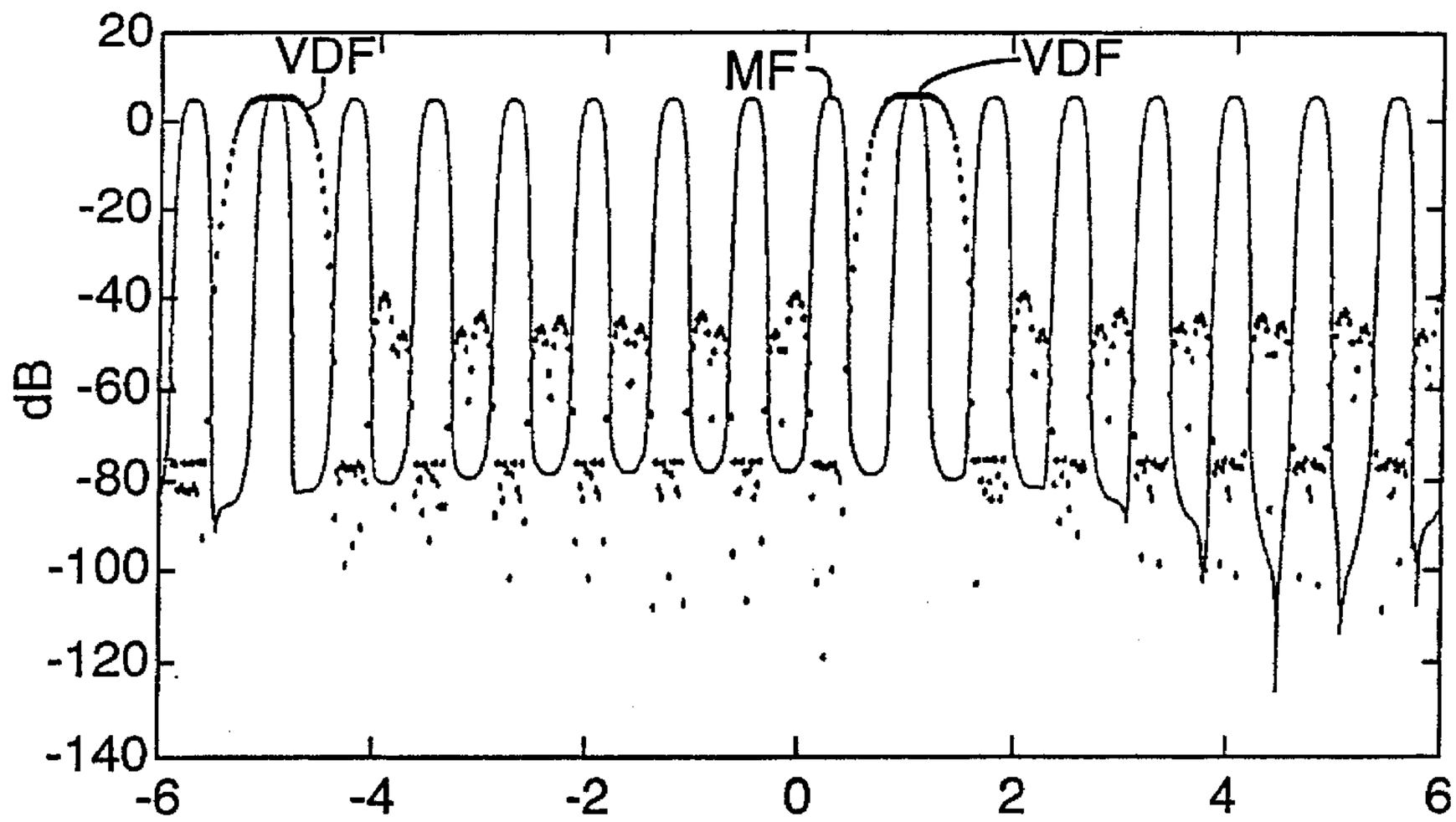
VDF
MF
VDF
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0
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dB
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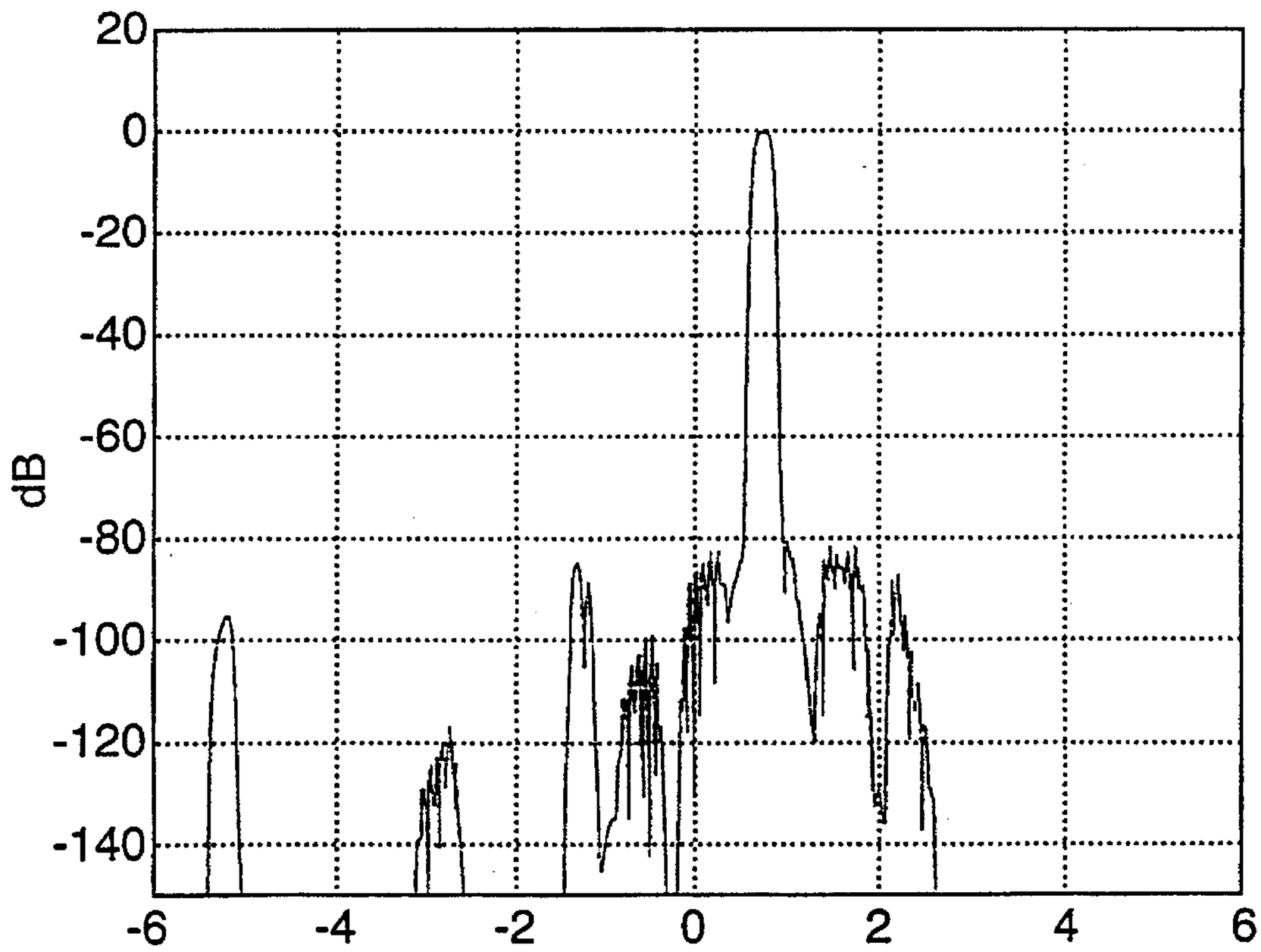
0
2
4
6

FIG. 20


20
0
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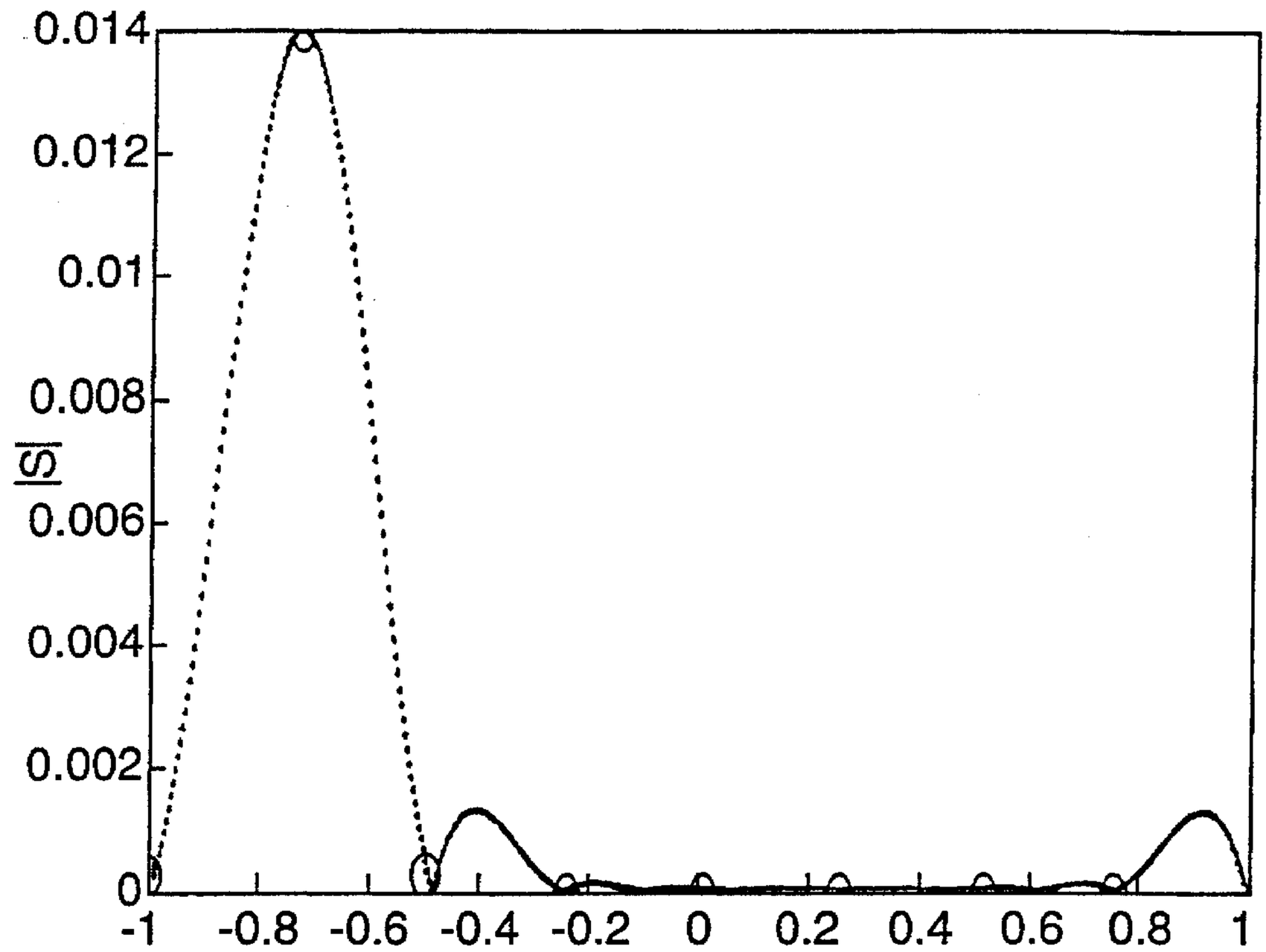
dB
-6
-4
-2
0
2
4
6

FIG. 21


0.014
0.012
0.01
0.008
|S|
0.006
0.004
0.002
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0.8
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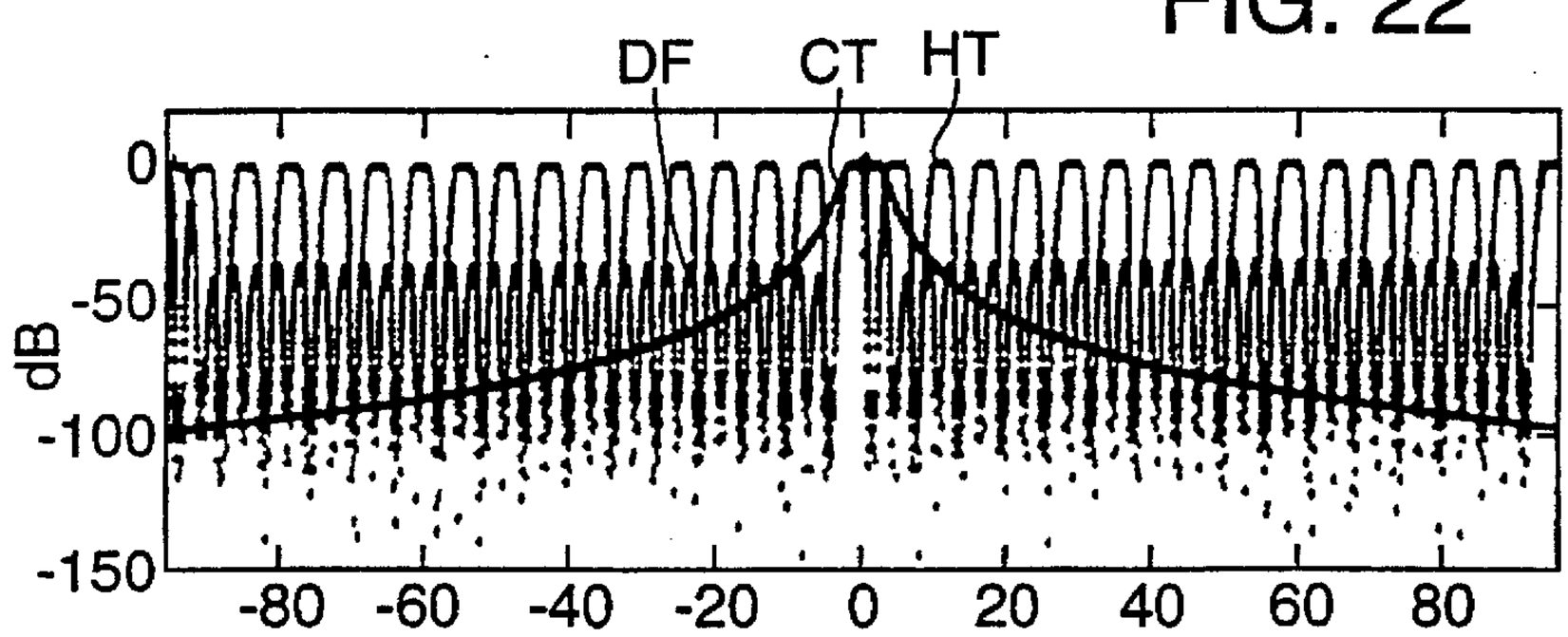
FIG. 22


DF
CT
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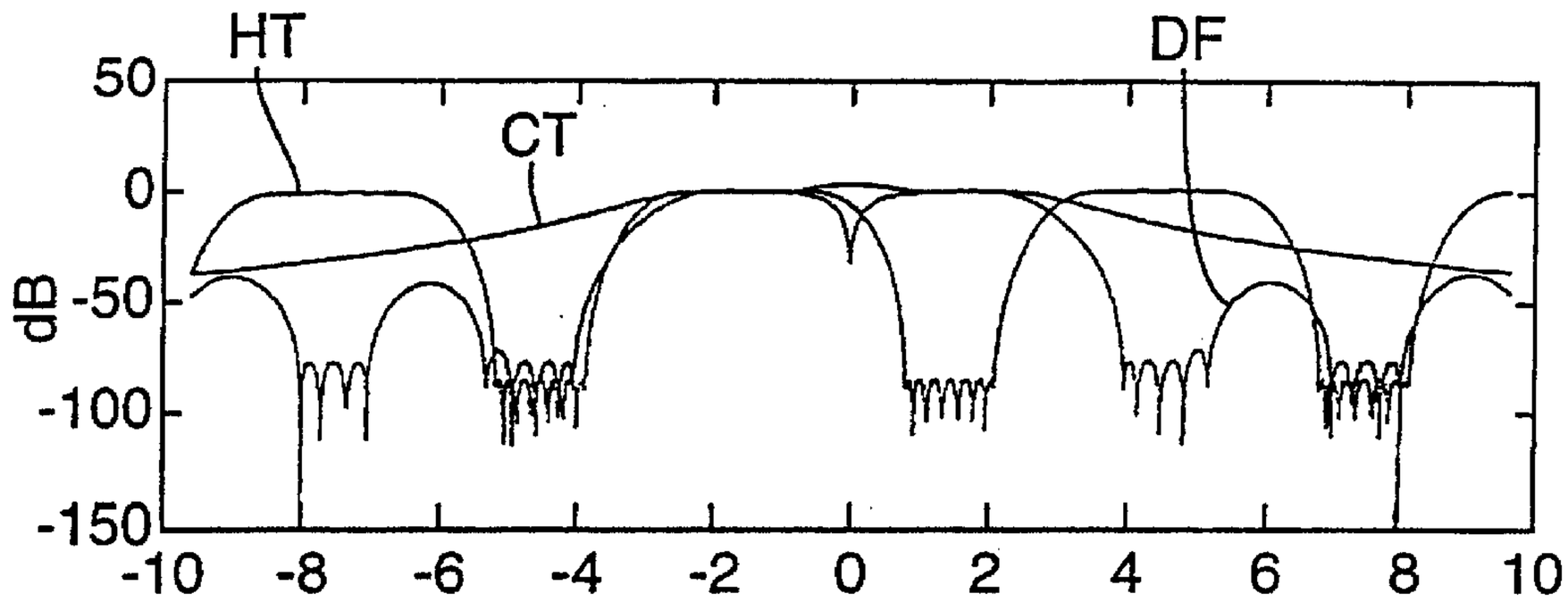
80

FIG. 23



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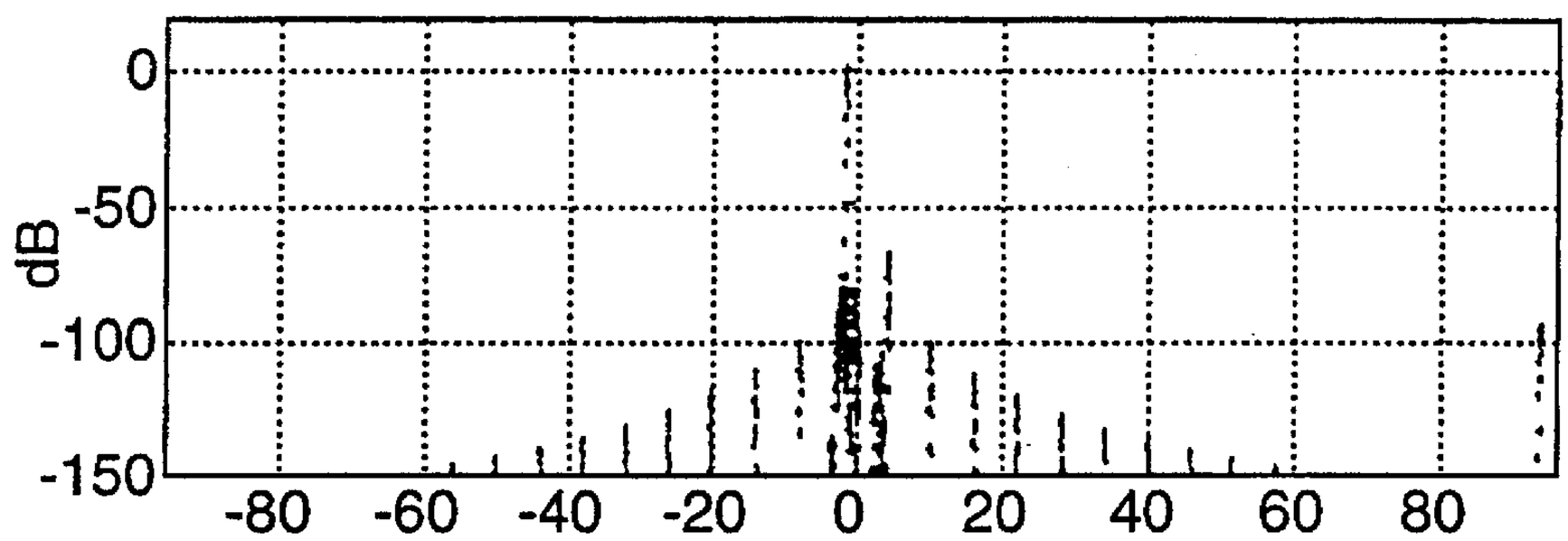


FIG. 25

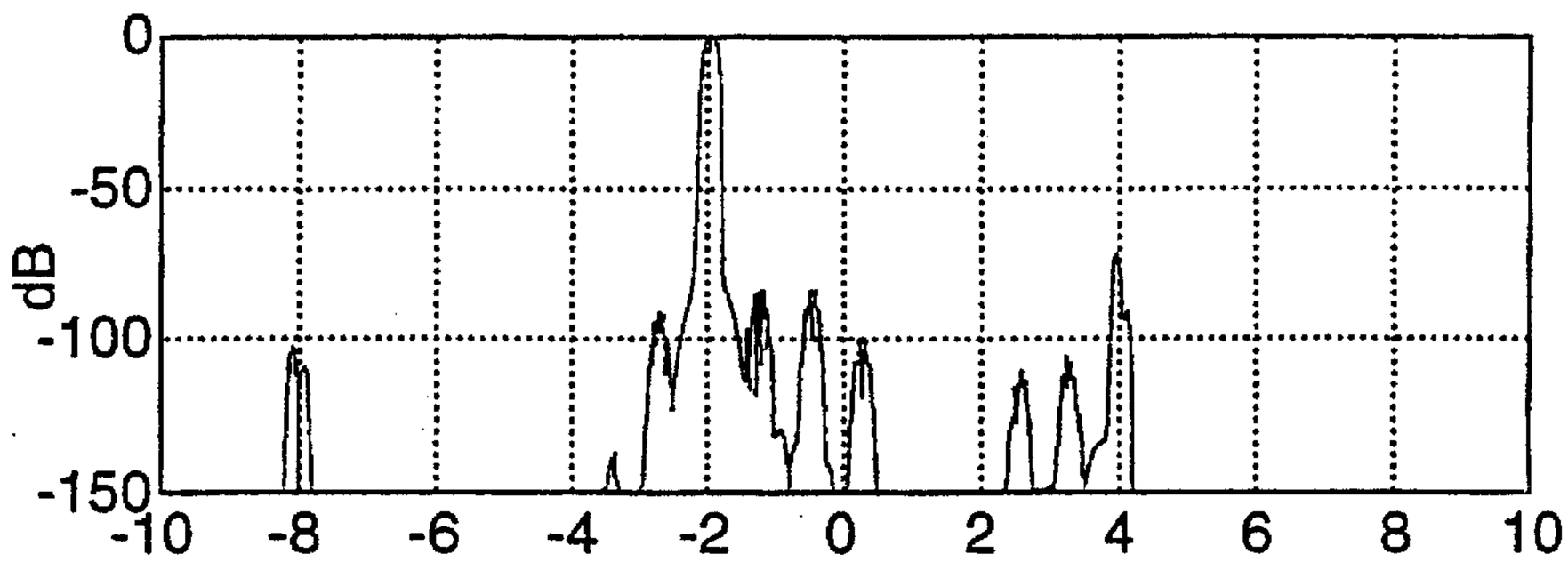
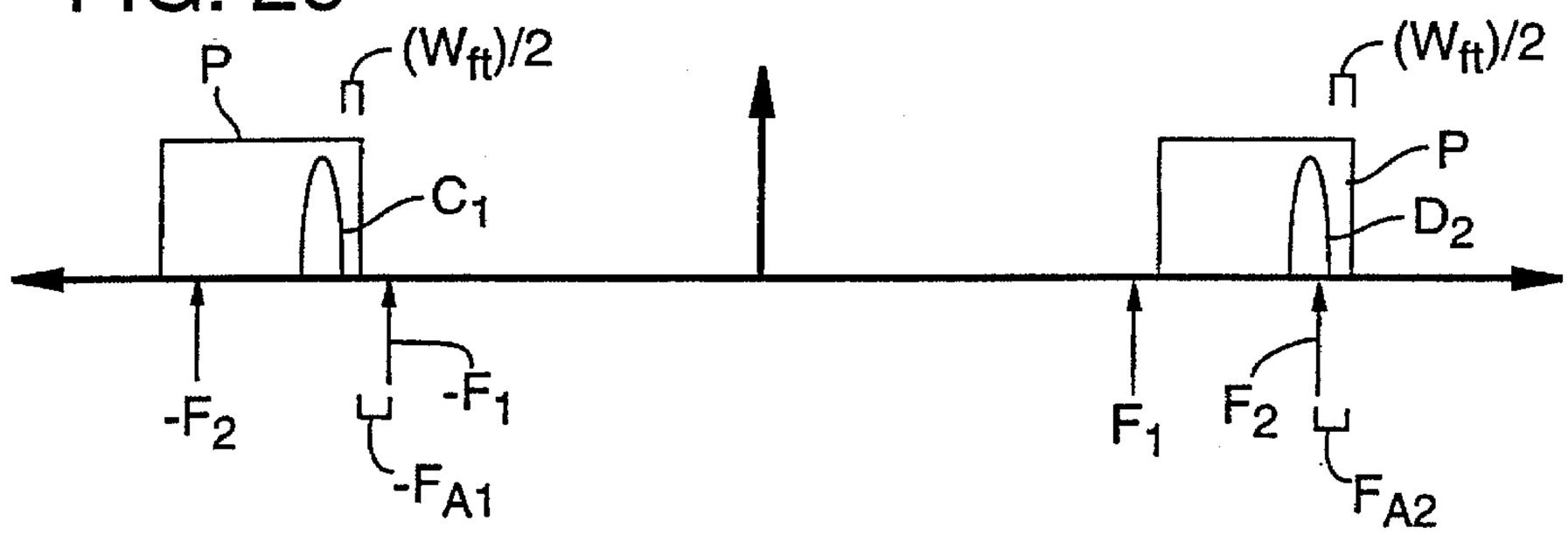


FIG. 26



SIMPLIFIED HIGH FREQUENCY TUNER AND TUNING METHOD

This application is a continuation of U.S. patent application Ser. No. 09/317,781 filed May 24, 1999, now U.S. Pat. No. 6,427,068 which is a divisional of U.S. patent application Ser. No. 08/713,761 filed Sep. 13, 1996 now U.S. Pat. No. 5,937,341.

FIELD OF THE INVENTION

This invention relates generally to devices and methods for receiving and transmitting RF signals. More particularly, this invention relates to an especially useful device and method for receiving and tuning RF signals, with quadrature mixing to a near baseband passband performed in continuous-time and image rejection and translation to baseband performed in discrete-time. The device may also be adapted to transmit RF signals if desired.

BACKGROUND OF THE INVENTION

Standard RF receiver design incorporates conversion of incoming high frequency signals to one or more intermediate frequencies, the last of which is then converted to baseband. A mixer and image rejection filter are required at each stage, resulting in complexity proportional to the number of stages. Such complexity is undesirable, particularly for mobile communications applications where size, power consumption, and cost per unit must be minimized.

Various approaches have been taken to reduce the size, power consumption, and cost of receivers. One approach is to perform nearly all of the receiver functions in the discrete-time domain in a DSP (digital signal processor) device. This results in high DSP performance requirements and cost. Other approaches employ discrete-time processing for baseband and for some intermediate frequency operations, reducing the DSP performance requirements, but still requiring at least one high performance continuous-time image rejection filter.

Direct conversion receivers offer a potential alternative for avoiding some of the limitations of other approaches. Receivers of this type employ quadrature mixing directly to baseband. Discrete-time processing can be efficiently utilized at baseband frequencies to demodulate the signal of interest, employing the quadrature baseband signals to utilize the entire signal spectrum centered at baseband. The complex-valued signal comprised of the I, Q samples allows the faithful representation of the signal of interest on both sides of baseband without distortion from images from opposite sides of baseband. Thus only a single continuous-time frequency conversion stage need be employed. No preselecting bandpass filter is required to eliminate an undesired mixing image, so that a broad tuning range is possible.

Despite the above potential advantages, direct conversion receivers also present problems including: (1) 1/f noise, which dominates active devices at low frequencies, particularly below 100 Hz, (2) time-varying DC offsets which can saturate the later stages of the baseband signal chain, (3) products of self-mixing of strong signals which can be present at baseband, (4) relatively small phase and amplitude errors between channels considerably reduce image rejection, and (5) fairly sharp anti-aliasing filters are required and can distort the desired signal if not carefully designed and precisely matched.

These problems are not unique to direct conversion receivers. An example of a receiver that converts to a non-zero intermediate frequency but remains vulnerable to the low-frequency problems listed above is illustrated in FIG. 13 of U.S. Pat. No. 5,875,212 to Fleek et al.

Attempts have been made to provide the advantages of direct conversion without the disadvantages by "notching out" DC from the baseband signal. This method performs well only if the signal type contains little or no information at or near DC. If the notch at DC is sufficiently narrow to minimize loss of information, the problems listed above related to amplification at or near DC are not eliminated.

Attempts have been made to avoid losing the information at and near DC and avoid the need for image rejection by translating a desired channel frequency from a channelized frequency spectrum to a frequency offset a small fixed amount from baseband, with the offset large enough to move the DC portion of the channelized signal into a passband which excludes DC, but small enough to prevent the next adjacent channel from appearing in the passband. This technique may preserve the DC portion of the signal, but requires sharp cut-off highpass and anti-aliasing filters and, because of the proximity of the passband to DC, still suffers somewhat from the other problems listed above.

Another known approach has been to perform image-rejection downconversion of an RF tuning range to a relatively wide intermediate frequency range with a local oscillator having no specified relationship to frequencies of RF channels within the tuning range. For example, W. Baumberger in "A Single-Chip Image Rejecting Receiver for the 2.44 GHz Band Using Commercial GaAs-MESFET-Technology" discloses the use of a 150-MHz intermediate frequency range (from 130 to 280 MHz) in a receiver having a tuning range on the order of 500 MHz. Another example is found in Published EPO Application 0 651 522 by M. Pesola, in which FIG. **3** illustrates the use of two radio frequency bands on opposite side of a local oscillator frequency, selected using either one of the outputs of a mixer attenuating the image frequency. Pesola also discloses the use of intermediate frequencies having widths of 100 kHz and 1 MHz, and a relatively high frequency of 100 MHz, with image-rejection downconversion using a channel-dependent local oscillator. This disclosed arrangement suffers from the inefficient use of a relatively high intermediate frequency (on the order of 100-1000 times the bandwith).

SUMMARY OF THE INVENTION

In accordance with the present invention, a high frequency spectrum of interest is translated in continuous-time to a near-baseband passband by quadrature mixing, preferably with a coarse-tuned local oscillator, producing I and Q signals in approximate quadrature relation. The I and Q signals are then filtered in continuous-time to remove DC and to prevent unwanted aliasing upon digital conversion, and are then converted to digital I and Q signals.

In digital processing, various steps are performed including (1) phase correction (optionally including group delay correction) and amplitude correction between the I and Q signals, (2) rejection of an image signal by means of complex filtering or a Hilbert transform pair and adder, (3) further bandlimiting, and, (4) translation of the desired signal from the near-baseband passband to baseband, which step may include digital fine-tuning over the near-baseband passband. If the desired signal is part of a channelized spectrum, the digital fine-tuning capability may be omitted or reduced to a coarse step-wise digital tuning capability with steps equal to the channel spacing, but a translation from near-baseband to baseband is still performed. These

steps may be performed in combination and in various orders to achieve the desired effect.

The inventive tuning method provides certain advantages of direct conversion receivers, including preferably a single continuous-time down-conversion stage, lack of image rejection filters with resulting wide possible tuning range, and relatively low frequency at conversion to discrete-time, allowing lower discrete-time processing rates and simplified decimation filter architecture. The inventive method also avoids the problems of 1/f noise and DC offset and self-mixing by avoiding the need for analog amplification of signal frequencies at baseband or only slightly offset from baseband.

The inventive tuning method further provides certain unique advantages.

For example, some significant advantages result from the inventive method's optimal division of tasks between continuous-time and discrete-time components.

In the inventive method, continuous-time components perform those tasks for which they are well suited, particularly the initial downconversion of a high frequency signal, while discrete-time components perform the tasks for which they are well suited, such as signal processing only at baseband and near baseband frequencies, yielding both relaxed continuous-time component tolerances and relatively reduced discrete-time processing speed and power requirements.

Further, the size and location of the near-baseband passband utilized in the invention and of the associated digital fine-tuning range or channelized spectrum channel spacing, if any, are so organized that the step size of the coarse-tuned local oscillator may be set to about twice the digital tuning range without any loss of spectrum coverage. The doubled step size relaxes the local oscillator requirements and reduces phase noise generated by the local oscillator. This relaxation of local oscillator (typically a PLL) requirements allows the local oscillator to cover a wider frequency range, so that the invention can take better advantage of the wide tuning range afforded by the lack of an image-rejection filter.

In the preferred embodiments, the invention also includes a type III Hilbert transform, i.e., a case III FIR phase-inverting allpass filter, for image rejection processing. The near-baseband passband utilized in the invention is then optimally sized and located for use with a type III Hilbert transform such that substantial computational and memory resource savings are realized while maintaining excellent performance.

A fuller appreciation of the above advantages and features and of additional advantages and features of the invention will be gained by those of skill in the art from the detailed description of the preferred embodiments which follows, and from practice of the invention itself.

BRIEF DESCRIPTION OF THE DRAWINGS

FIG. **1** is a diagram of a device according to the present invention.

FIG. **2** is a diagram of the near-baseband passband utilized by present invention.

FIGS. **3** and **4** are diagrams illustrating the use of the near-baseband passband of the present invention with channelized frequency spectra.

FIGS. **5** and **6** are diagrams illustrating the doubled local oscillator step size achievable according to the present invention.

FIG. **7** is a diagram showing the preferred size and location of the near-baseband passband of the present invention in relation to various characteristics of various preferred elements of the present invention.

FIG. **8** is a diagram of presently preferred embodiments of the present invention.

FIG. **9** is a diagram showing certain aliasing regions of the near-baseband passband together with highpass frequency response curves for filters for use in the present invention.

FIG. **10** is a diagram showing additional aliasing regions of the near-baseband passband according to an embodiment of the present invention.

FIG. **11** is a diagram of a preferred embodiment of a decimating filter for use in an embodiment of the present invention.

FIG. **12** is a simulated frequency response curve of the filter of FIG. **11**.

FIG. **13** is the simulated frequency response curve of FIG. **12** shown on a smaller scale.

FIG. **14** is a simulation plot of quantization noise both with and without aliased quantization noise.

FIG. **15** is a diagram illustrating the operation of a Hilbert transform modified according to the present invention.

FIG. **16** is a diagram illustrating the presently preferred method of correcting phase errors used in the present invention.

FIG. **17** is a diagram illustrating the generation of coefficients for a variable group-delay allpass filter usable in an embodiment of the present invention.

FIGS. **18** and **19** are simulated frequency response curves of certain elements of an embodiment of the present invention.

FIG. **20** is a simulated frequency response curve of an embodiment of the present invention.

FIG. **21** is a simulated envelope detector output of an embodiment of the present invention.

FIGS. **22** and **23** are simulated frequency response curves of certain elements of an embodiment of the present invention.

FIGS. **24** and **25** are simulated frequency response curves of an embodiment of the present invention.

FIG. **26** is an additional diagram illustrating the use of the near-baseband passband with channelized frequency spectra.

DETAILED DESCRIPTION OF A PREFERRED EMBODIMENT

A significant aspect of the present invention is the basic division of functions between discrete-time (digital) and continuous-time (analog) components characteristic of the invention, which is described with reference to FIG. **1**.

In FIG. **1**, analog portion **12** of a device **10** according to the present invention receives an incoming signal from a preferably removable/exchangeable antenna **16**. A suitable broad-band or tunable RF amplifier **18** then amplifies the signal. Alternatively, the invention could also be used to tune an intermediate frequency from previous analog processing, rather than an incoming signal directly from antenna **16** and amplifier **18**.

The signal is then split into two signal paths and fed to first and second mixers **20**, **22**. The first and second mixers **20**, **22** are supplied with a quadrature local oscillator signal from a preferably coarse-stepped local oscillator **24**. The mixing operation translates to a near-baseband passband an upper high frequency spectrum of interest from above the

frequency F_{LO} of the local oscillator **24** and a lower high frequency spectrum of interest from below the frequency F_{LO} of the local oscillator **24**, producing I and Q signals in approximate quadrature relation.

The near-baseband passband is sufficiently low to provide substantial efficiency gains in the subsequent digital processing, but does not include baseband. The near-baseband passband is also sufficiently high to allow a fairly relaxed transition from a cutoff at or near DC to the lower edge of the passband. Problems such as self-mixing products and DC offsets and 1/f noise are avoided by high-pass filtering with a relaxed transition band in filters **26** and **28**. Unwanted aliasing is prevented by low-pass filtering in filters **26** and **28**. The I and Q outputs from the analog portion **16** of the device **10** are then passed to the digital portion **14** of the device **10**.

At least three operations are performed within the digital portion **14** of the device **10**. First is analog to digital conversion. The I and Q signals are converted individually into digital signals. Second, phase errors (optionally including group delay errors) and amplitude errors between the I and Q channels are corrected, particularly to maximize image rejection at and/or near the frequency of the desired signal, and the channels are combined by a Hilbert transform pair and summing, or filtering with complex coefficients is employed, in order to reject the undesired mixing image, particularly at the frequency of the desired signal. Third, a portion of the now image-rejected signal containing the desired signal is translated to baseband. The second two operations may be performed in various orders or to some extent simultaneously according to the particular design and programming of the digital portion **14** of the inventive device **10**.

The above-described division of functions into analog and digital domains, together with the use of the properly located near-baseband passband, provides important advantages. The number of analog components is minimized, and the analog components are employed for those tasks to which they are most suited: the conversion of high frequencies to low frequencies. The digital processing is used only at lower frequencies, allowing lower sampling rates and quantization resolutions to be employed without substantial loss of signal characteristics, resulting in decreased memory, processing power, and electrical power requirements. The near-baseband passband avoids analog processing of signals at or close to DC, thereby avoiding or substantially diminishing problems associated with 1/f noise and DC offsets and self-mixing products. Use of quadrature mixing with subsequent image rejection avoids the need for relatively high-performance image rejection filters in the analog portion. Correction in digital processing of phase and amplitude deviations from the desired quadrature relation allows relaxation of otherwise relatively strict matching and performance requirements for the analog filters. Fine-tuning, if employed, is preferably performed in the digital domain, leaving only coarse-tuning by a coarse-stepped local oscillator to be performed in analog processing, thereby reducing the complexity of the local oscillator and the generation of phase noise.

Another significant aspect of the present invention is that the local oscillator can be a coarse-tunable local oscillator having a step size S that is twice as large as would typically be permitted, given the range of the digital fine-tuning employed, or given the channel spacing of the channelized spectrum and the digital channel tuning employed. This is achieved by proper positioning and sizing of the near-baseband passband and the tuning range or channelized tuning/translation range of the digital tuning process, and takes advantage of the fact that complex I, Q signals contain twice the spectral information of a real signal.

As illustrated in FIG. **2**, the near-baseband passband P may be defined with reference to a lower frequency F_1 and an upper frequency F_2. To achieve the preferred effective doubling of the local oscillator step size S, F_1 and F_2 are chosen such that $F_1=k\cdot(F_2-F_1)$, where k is a positive integer, and S is set to $2\cdot(F_2-F_1)$. This insures that the center frequency of any desired incoming signal can be translated to within the positive frequency range of F_1 to F_2 inclusive or the negative frequency range of $-F_1$ to $-F_2$ inclusive by mixing with the appropriate local oscillator frequency. The use of complex I, Q signals allows the positive frequency range to be distinguished from the negative frequency range.

For embodiments of the invention designed to tune essentially any desired frequency from within a given RF range, the digital tuning process employed preferably has a range extending from F_1-F_H to F_2+F_H, where F_H is an appropriate hysteresis amount greater than or equal to zero, the effects and usefulness of which will be explained hereafter. The near-baseband passband P is then defined so as to extend from F_1-F_A to F_2+F_A as shown in FIG. **2**, where F_A is a frequency adjustment equal to at least about $W/2+F_H$, where W is the maximum bandwidth of the desired signals to be received. This ensures that all of the bandwidth of any signal having a center frequency tunable by the digital tuning process will fall within the near-baseband passband.

A device of the present invention designed to tune essentially any frequency can of course be utilized to receive channelized signals. If decreased digital processing is desired in an embodiment designed for channelized signal reception, the full digital tuning capability over the entire near-baseband passband can be restricted to discrete digital tuning in the form of either (1) a translation to baseband from a chosen frequency within the near-baseband passband (preferably the midpoint between F_1 and F_2), or (2) a step-wise tuning of selected channelized frequencies from the near-baseband passband to baseband. A small amount of fine-tuning may be retained if desired for fine-tuning around the discrete channelized frequency(ies) within the near-baseband passband.

In embodiments of the present invention employing discrete digital tuning, the center frequency of each channel of any given channelized spectrum will be translated to within the frequency range from F_1 to F_2 inclusive or from $-F_1$ to $-F_2$ inclusive by mixing with one of the various possible local oscillator frequencies.

One possibility for selecting F_1 and F_2, with reference to which the near-baseband passband may be defined, is choosing F_1 and F_2 such that $F_2-F_1=N\cdot C$ where C is the channel spacing and N is the number of channels to be contained within the near-baseband passband. F_1 and F_2 may be further chosen, along with the local oscillator frequency, such that F_1 and F_2 each fall at the midpoint between adjacent channels after translation of the channel frequencies by mixing with the local oscillator signal. This is possible where the permissible local oscillator frequencies are at frequencies one-half of a channel spacing C displaced from integer multiples of the channel spacing, and is illustrated for the case N=1 in FIG. **3**.

In FIG. **3**, adjacent channels of bandwidth W have been down-converted by mixing with the local oscillator signal. F_1 and F_2 fall between adjacent down-converted channels. The near-baseband passband P extends from F_1-F_A to F_2+F_A, where F_A is a frequency adjustment equal to ½(W–

$C+W_{ft}$), where W_{ft} is the width of the digital fine-tuning, if any, provided for fine-tuning around each channel.

FIG. **4** shows the near-baseband passband P in a channelized embodiment using with N=3, W=C, W_{ft}=0, and F_H=0. The local oscillator frequency is in this case an integer multiple of the channel spacing. The near-baseband passband P is defined with reference to F_1 and F_2 and the full-range digital tuning frequency adjustment F_A, which is equal to $(½)\cdot(W+W_{ft})$. F_2-F_1 in this case equals $(N-1)\cdot C$.

The decreased digital processing realized with channelized operation may be obtained even with irregular intervals between channels or with local oscillator step sizes not evenly divisible by the channel spacing or with channels not located at integer multiples of the channel spacing. A generalized case is illustrated in FIG. **26**. For a given channelized spectrum and a given set of possible local oscillator frequencies, the center frequency of every channel is translatable by one of the local oscillator frequencies to within the range F_1 to F_2 inclusive or the range $-F_1$ to $-F_2$ inclusive. The channel having a center frequency after such translation the absolute value of which is closest to (but not greater than) F_2 is used to determine the upper edge of the near-baseband passband P. The center frequency of such channel after translation may be either negative or positive, but is illustrated in FIG. **26** as positive channel C_2. The upper edge of the near-baseband passband P (the edge with the greatest absolute value) is then located at the edge of the bandwidth of channel C_2 furthest from DC, plus another $W_{ft}/2$ from DC, which is half the width of any digital fine tuning range. Similarly, the channel having a center frequency after translation the absolute value of which is closest to (but not less than) F_1 is used to determine the lower edge of the passband. The center frequency of such channel after translation may be either negative or positive, but is illustrated in FIG. **26** as negative channel C_1. The lower edge of the near-baseband passband P (the edge with the least absolute value) is then located at the edge of the bandwidth of channel C_1 closest to DC, plus another $W_{ft}/2$ toward DC. The resulting frequency adjustments F_{A1} and F_{A2} are shown in FIG. **26**.

Note that in the above examples and throughout the specification and claims, it should be understood that the near-baseband passband refers to that portion of the frequency spectrum to which signals of interest are to be translated in continuous time for further processing in discrete time. The actual physical passband created by the frequency response of the filters **26** and **28** of FIG. **1** may, of course, be larger than the near-baseband passband itself, and indeed it is preferred that the actual passband be somewhat larger, so that the corners of filters **26** and **28** do not appear within the near-baseband passband. Preventing the corners from appearing in the near-baseband passband reduces group delay variation that can cause degradation in image rejection and can worsen intersymbol interference.

Whether in an embodiment with essentially continuous digital fine-tuning or with channelized digital tuning, the frequencies F_1 and F_2 are selected such that $F_1=k\cdot(F_2-F_1)$, where k is a positive integer. Most preferred is k=1 as in FIG. **2**, but other values can be used, such as k=2 as shown in FIG. **4**. The effective doubling of the permissible step size S of the local oscillator results in part from utilization of this equation as illustrated below with respect to FIGS. **5** and **6**.

For a given local oscillator frequency F_{LO}, both an upper high frequency spectrum of interest and a lower high frequency spectrum of interest are translated to the near-baseband passband. By means of image-rejection processing employed in the digital domain, either the near-baseband image of the upper high frequency spectrum of interest or the near-baseband image of the lower high frequency spectrum of interest may be rejected, allowing selection of either the upper high frequency spectrum of interest or the lower high frequency spectrum of interest for further processing. Because of the positioning and size of the near-baseband passband and associated digital tuning, whether continuous fine-tuning or stepwise tuning, alternate selection of the upper and lower high frequency spectra of interest can be used to provide non-redundant coverage of the broadband frequency spectrum from which the desired signal is to be received. Any desired frequency may then be translated to the near-baseband passband with the local oscillator step size S set to twice the digital tuning range, i.e., S set equal to $2\cdot(F_2-F_1)$.

FIG. **5** shows a portion of the positive frequency spectrum graphed on a linear scale. Each possible value of F_{LO} is indicated with an arrow and labeled with a letter or letters and a backslash. The letters are also used to label a number of frequency regions with a channelized signal in each. This represents an embodiment for a channelized spectrum with a near-baseband passband sized to fit one channel, as in FIG. **3**. The letters labeling each possible value of F_{LO} correspond to the letters labeling the regions translated to the near-baseband passband by that value of F_{LO}. If F_{LO} is at A\D, for example, regions A and D are translated to the near-baseband passband. The letter to the left of the backslash corresponds to the letter labeling the lower high frequency spectrum of interest for a given F_{LO}, while the letter to the right corresponds to the letter labeling the upper high frequency spectrum of interest for that given F_{LO}.

To translate to the near base-band passband the channelized signal frequency within region F, for example, the local oscillator frequency F_{LO} would be set to the C\F position. The desired signal frequency would then fall within the upper high frequency spectrum of interest of the local oscillator frequency. The image of region C, the lower high frequency spectrum of interest, would be rejected by the digital image rejection processing, leaving F as the selected region.

Similarly, to tune the channelized signal frequency within region G shown in FIG. **5**, the local oscillator frequency F_{LO} would be set to the G\J position. The desired signal frequency would then fall within the lower high frequency spectrum of interest of the local oscillator frequency, and the image of region J would be rejected by the digital image rejection processing, leaving G selected.

While k=1 is preferred as noted above, other values are possible such as k=2, illustrated in FIG. **6** for an embodiment with essentially continuous digital fine-tuning. The upper and lower high frequency spectra of interest are now each about two tuning ranges separated from the applicable local oscillator frequency F_{LO}, but the same complete coverage, with a step size $S=2\cdot(F_2-F_1)$, is provided. To tune a signal frequency within region C, for example, the local oscillator frequency would be set to the C\H position, and the lower high frequency spectrum of interest would be selected.

The proper F_{LO} to receive a given desired signal frequency F_t may be found by any appropriate method. For example, the proper F_{LO} may be found generally by setting $NLO=floor(F_t/S+½)$, which is the factor NLO such that $NLO\cdot S$ is the nearest F_{LO} to the desired signal frequency F_t. If $NLO\cdot S \geqq F_t$, the proper F_{LO} to employ to translate F_t to the near-baseband passband is then given generally by $(NLO+(-1)^k\cdot floor(k/2+½))\cdot S$, with F_t found in the upper high frequency spectrum of interest if $(-1)^k<0$, and in the lower high

frequency spectrum of interest otherwise. Similarly, if $NLO \cdot S \leqq F_t$, the proper F_{LO} is given by $(NLO-(-1)^k \cdot \text{floor}(k/2+½)) \cdot S$, with F_t in the upper high frequency spectrum of interest if $(-1)^k>0$, and in the lower high frequency spectrum of interest otherwise. (The ambiguity in F_{LO} selection at $NLO \cdot S=F_t$ is caused by the overlap of adjacent high frequency spectra of interest as seen in FIG. **6**, which allows a desired signal of frequency F_t equal to $NLO \cdot S$ to be translated to the near-baseband passband by either of two possible values of F_{LO}.)

To avoid having to repeatedly change the value of NLO when fine-tuning around a signal frequency F_t about equal to $NLO \cdot S$, the hysteresis amount F_H for essentially continuous fine-tuning embodiments may be set to a value greater than zero. This allows fine tuning on both sides of a signal frequency F_t equal to about $NLO \cdot S$ with only one local oscillator frequency F_{LO}, and widens each high frequency spectrum of interest by $2 \cdot F_H$. If a desired frequency F_t moves out of a first widened high frequency spectrum of interest associated with a first F_{LO}, a second F_{LO} is selected. If F_t then moves back into the first widened high frequency spectrum of interest, the second F_{LO} is maintained until F_t is within the first widened high frequency spectrum of interest by the distance F_H. Thus excessive switching from one F_{LO} to another is prevented.

If the invention is used to tune an intermediate frequency from previous analog processing, either a local oscillator in the previous analog processing or the quadrature local oscillator **24** may be varied and the other local oscillator set to a fixed frequency. The effective doubling of the permissible step size S of the varied local oscillator is retained in either case through the alternate selection of the upper and lower high-frequency spectra of interest. Alternatively, the invention can tune an intermediate frequency signal produced by analog mixing without requiring a coarse-tunable local oscillator.

The near-baseband passband can include a frequency range extending from a lower frequency F_1-F_A to an upper frequency F_2 F_A, where F_A is a given frequency adjustment. While the actual frequencies F_1 and F_2, selected as described above, will vary with the particular application, it is currently preferred that F_1 be at least about W, where W is the bandwidth of the channels to be received or the bandwidth of channelized signals to be received from a channelized spectrum. Also preferred is that, in general, F_2 be not more than about 150 kHz. In some applications, however, F_2 may be many times this generalized upper limit. It is currently preferred that F_2-F_1 be at least about 20 kHz. In embodiments with full-range digital tuning, it is currently preferred that F_2-F_1 be within the range of about $3 \cdot W$ to about $5 \cdot W$. The frequency adjustment F_A can be equal to, for example, the quantity $(½) \cdot (W-C+W_{ft})$, the quantity W/2, or the quantity $W/2+F_H$, where C is the channel spacing of a channelized spectrum and w_{ft} is of the width of any fine-tuning performed.

Another significant aspect of the present invention is the preferred use of a type-III Hilbert transform in the image rejection processing in the digital domain. A type III Hilbert transform enjoys nearly a 2:1 efficiency advantage over a similar standard type IV Hilbert transform, because every other impulse response sample is zero. The performance envelope of the type III Hilbert transform is symmetrical and centered on $f_s/4$ (where f_s is the sampling frequency employed); falling off symmetrically approaching DC and $f_s/2$. While the performance of the type III falls off relative to the type IV as frequencies approach the Nyquist frequency of $f_s/2$, the present invention avoids any disadvantage from this characteristic as will be seen below with respect to FIG. **7**.

Due to the preferred spacing from DC of the near-baseband passband of the present invention as illustrated for example in FIG. **2**, the near-baseband passband is sufficiently separated from DC for an efficient and accurate Hilbert transform to be performed. The relatively wide transition band to DC also affords relaxed filter specifications. To take advantage of the type III transform's efficiency and to provide even more relaxed filter specifications, the present invention preferably employs a type III Hilbert transform with a sampling rate R entering the Hilbert transform equal to $2 \cdot (F_1+F_2)$. This is equivalent to centering the near-baseband passband at R/4. Some of the advantages of this arrangement are illustrated for the case k=1 in FIG. **7**.

FIG. **7** shows the near-baseband passband P of the present invention located to encompass a digital tuning range between F_1 and F_2 with $F_1=k \cdot (F_2-F_1)$ with k being a positive integer, in this case 1. The step size S of the analog local oscillator is also shown for reference. The use of a type III Hilbert transform with an entering sampling frequency $R=2 \cdot (F_1+F_2)$ results in the illustrated performance curve H for the preferred type III Hilbert transform. The performance curve H is symmetrical with the best performance at the location of the near-baseband passband P, with symmetrically reduced performance toward DC and the Nyquist frequency of R/2.

The near-baseband passband is also situated so as to substantially avoid 1/f noise represented by the $N_{1/f}$ spectrum shown, and quantization noise Q from a presently most preferred delta-sigma analog to digital conversion. The transition bands T are also sufficiently broad to relax filtering requirements in both the analog and digital domains as will be shown in greater detail below.

A pair of preferred embodiments of the device of the present invention implementing the methods and having the characteristics and advantages discussed above are shown in greater detail in FIG. **8**. The device **10** shown in FIG. **8** corresponds to the device **10** shown in FIG. **1**, but with details of presently preferred embodiments shown in the digital portion **14** of FIG. **8**. Accordingly, analog portion **12** shown in FIG. **8** is as described above with reference to FIG. **1**.

The analog I and Q signals are received into digital portion **14** from the analog portion **12** of the device **10** and are converted into digital signals by delta-sigma modulators **30**, **32** most preferably third-order delta-sigma modulators, with one-bit wide output. The delta-sigma modulators sample the I and Q signals at an over-sampling rate R_O. Decimation filters **34** and **36** filter the output of the delta-sigma modulators so as to substantially reject frequencies which would alias into the near-baseband passband on decimation, and decimate the signal, such that the output sample rate is equal to R, the desired input sampling frequency at a Hilbert transform pair, comprised of a Hilbert transform **38** and allpass filter **40**, which follows.

An alternate embodiment, shown in FIG. **8** by the dashed-line alternate signal paths I_A and Q_A, does not employ oversampling. Instead, the I and Q signals are sampled by analog to digital converters **42** and **44** at the rate R, the input sampling frequency at the Hilbert transform pair **38** and **40**, and converted into digital signals preferably 12 to 16 bits wide, depending on the dynamic range requirements of the application. Thus no decimation is required between analog to digital converters **42** and **44** and Hilbert transform pair **38**,

40. For this alternate embodiment, the near-baseband passband of the present invention provides somewhat relaxed anti-aliasing lowpass filter specifications, as illustrated in FIG. 9.

The near-baseband passband P of the present invention, for the case k=1, is shown in FIG. 9. P ends at $(k+1)\cdot S/2+F_A$. The region AR_R is the first (lowest frequency) region to alias into P at a sampling rate of R. AR_R begins at $R-(k+1)\cdot S/2-F_A$. Accordingly, the passband of the anti-aliasing filter represented by response curve **46** must extend at least to $(k+1)\cdot S/2+F_A$, while the stop band must begin at or before $R-(k+1)\cdot S/2-F_A$. This prevents aliasing into P while allowing a fairly relaxed transition band between the passband and stop band. A highpass filter would preferably be employed with a passband beginning at the lower edge of P, which is given by $k\cdot S/2-F_A$. These two filters together then comprise filter **26**, for example, in FIG. **8**. The lowpass filter can be an eighth order switched-capacitor elliptical lowpass filter, for example. The passband of the filter **26** preferably extends even beyond the edges of P such that the corners of the filter, with their typically large group delay, are not within P.

The relatively high order low-pass filters typically needed for moderately relaxed transition bands such as the transition band in response curve **46** of FIG. **9** can cause less efficient image rejection due to small variations in pole and zero locations between the filters **26** and **28** in the device **10** of FIG. **8**. In the most preferred embodiment shown in FIG. **8** by the I and Q solid line signal paths, oversampling allows use of much lower order anti-aliasing filters, with corresponding improvements in image rejection.

The relaxed anti-aliasing filter transition band obtainable with oversampling is illustrated in FIG. **9**, where R_O is the oversampling sampling rate, with $R_O=M\cdot R$ where M is the rate of oversampling. M=3 is shown for illustration purposes in FIG. **9**. AR_{Ro} is then the first aliasing region, i.e., the lowest frequency region to alias into P at a sampling rate of R_O. AR_{Ro} begins at $R_o-(k+1)S/2-F_A$. Accordingly, the stopband of the anti-aliasing filter represented by response curve **48** must begin at or before $R_O-(k+1)\cdot S/2-F_A$, with the same passband region as response curve **46**. Response curve **48**, together with oversampling, thus prevents aliasing into P while allowing a very relaxed transition band between the passband and stop band of the anti-aliasing filter. In practice, even greater oversampling rates than 3 are desirable, with M=32 currently most preferred. A 2-pole Chebychev type I low pass filter is then preferred for the low pass filter portion of filters **26** and **28**.

Note that, as discussed with reference to FIG. **26**, the frequency adjustment F_A may have differing values at the upper and lower edges of the near-baseband passband P.

In the most preferred embodiment, decimating filters **34** and **36** follow the delta-sigma modulators **30** and **32**, respectively. One of the decimating filters **34** and **36** preferably includes a group delay correction. The output of the delta-sigma modulators **30** and **32** is preferably one bit wide, allowing group delay correction to be implemented with a variable shift register in the signal path. One-bit signal width, together with the near-baseband passband and other features of the present invention, also makes practical an efficient single stage implementation of filters **34** and **36** with no multiplication required.

The aliasing regions of concern in the design of filters **34** and **36** are illustrated for example in FIG. **10**. The output of filters **34** and **36** is to be sampled at a rate R which is equal to R_O/M, where R_O is the over-sampling sampling rate. M=3 is used in FIG. **10** for illustration purposes. The first aliasing region AR_R is the first, i.e. lowest frequency, region to alias into P due to the sampling at rate R. Subsequent aliasing regions AR are also shown in FIG. **10** to the right of AR_R. The desired decimating filter should thus have a passband at P and stopbands of at least the same width as P at each aliasing region. All other frequency regions may be left unconstrained in the filter design process. Leaving regions which do not alias into the near-baseband passband P unconstrained, particularly with the size and position of P relative to R as preferred in the present invention, allows significant reduction in filter order and/or length such that a single stage decimation filter with good performance and reasonably low processing and memory requirements can be implemented.

A preferred design for a single stage decimation filter for use as filter **34** and/or **36** is shown in FIG. **11**. The filter includes: 32-bit registers **50***a*–**50***i*, operators **52***a*–**52***d*, look-up tables **54***a*–**54***d* with a look-up table address generator **56**, a 22-bit accumulator **58**, and a truncator **60** and a decimator **62**.

The 32-bit register **50***a* receives signal bits from the associated upstream one-bit delta-sigma modulator until the register **50***a* is full. Each time register **50***a* is full, the contents of each of the 32-bit registers **50***a*–**50***h* are shifted into the 32-bit registers **50***b*–**50***i* respectively. Because of the one-bit signal width and the symmetrical filter coefficients and folded-over filter architecture, the operators **52***a–d* can be used to efficiently determine the filter output without multiplication.

In the operators **52***a–d*, each of the 32 signal bits from one associated register is exclusive or-ed with the signal bit at the same distance from the center of the filter's delay line from the other respectively associated register. If the bits are not equal, a zero is added to the contents of the accumulator **58**. If the bits are equal and positive, twice the value of the applicable coefficient is added to the contents of the accumulator **58**, and if the bits are equal and negative, negative twice the value of the applicable coefficient is added to the contents of the accumulator **58**. This can be easily implemented by storing not the actual filter coefficient values in the look-up tables **54***a*–**54***d*, but twice the actual coefficient values. Then if two signal bits compared in the operator **52***a* for example are equal, the sign of one can be used as the sign bit to be sent to the accumulator **58** along with the doubled coefficient value from the look-up table **54***a*.

Each of operators **52***a*–**52***d* operates in parallel on two of the 32-bit registers **50***b*–**50***i*, and sends its output to accumulator **58** in parallel. Once all coefficients have been summed in accumulator **58**, truncator **60** takes only the 16 most significant bits from accumulator **58**. Decimator **62** represents the decimation performed in the operation of this filter.

Table I below contains Matlab® code for generating filter coefficients for the filter illustrated in FIG. **11** for use with the near-baseband passband of the present invention with k=1 and the near-baseband passband centered at ¼ of the Nyquist frequency. (Matlab® is software for digital signal processing analysis and development available from The MathWorks, Inc., Natick, Mass., U.S.A.) In the code in Table I, the variable fs is the sampling frequency entering the Hilbert transform and is set to 128 kHz. N is the filter order, which is set to 255, resulting in 256 FIR filter taps, which number is desirable as a power of 2 giving easier implementation in a DSP, and because it provides sufficient taps to significantly reduce noise at each alias of the near-

baseband passband. R is the weight of the stopband constraint versus passband constraint, and is set to 100, resulting in very high stopband rejection at the expense of some passband ripple. Mds is the over-sampling and decimation ratio, and is set to 32.

Table II below gives an example of doubled filter coefficient values for use in the look-up tables **54***a*–**54***d*. Note that look-up tables **54***a* and **54***b* require 32×13 bits of storage, while look-up table **54***c* needs 32×14, and look-up table **54***d* needs 32×16.

The resulting simulated frequency response of the filter in FIG. **11** is shown in FIGS. **12** and **13**. In FIG. **12**, the filter response at the near-baseband passband, located at 21.33–42.67 kHz in this example, may be seen. In FIG. **13** with a smaller scale, the repeating stopbands at aliasing regions of the near-baseband baseband passband may be seen, including stopbands at about 80–100 kHz, 140–160 kHz, etc.

One important criteria for judging the performance of the filter of FIG. **11** is the reduction of quantization noise from the delta-sigma modulator, particularly from higher aliasing regions which would alias into the near-baseband passband. FIG. **14** shows the noise due to quantization after filtering and decimation without (lower trace) and with (upper trace) the aliased quantization noise. Noise floors in the near-baseband passband located in this case at 21.33–42.67 kHz are still at quite acceptable levels, even with the addition of the aliased noise.

The preferred design of the Hilbert transform **38** of the device **10** in FIG. **8** is of course type III, but with at least one modification. Type III Hilbert transforms have an odd number of taps, with the center tap, and taps displaced an even number of taps from the center tap, set to zero. The Hilbert transform **38** is modified by having a variable non-zero coefficient present at the center tap, i.e., at the sample of its impulse response the index of which corresponds to half the length of the transform delay line. This modification enables the Hilbert transform **38** to function as if in parallel with an all pass filter with variable gain, as shown schematically in FIG. **15**. As the contribution from the allpass portion increases or decreases from zero, the phase change caused by the Hilbert transform is varied up or down from 90 degrees, allowing efficient correction of phase errors. The other coefficients of the Hilbert Transform may also be varied along with the central coefficient to implement correction of amplitude errors between the I and Q channels. Variable gain for amplitude error correction may also be implemented in the allpass filter **40** if desired.

In implementing any type of error correction between the I and Q channels, the errors should be corrected not to maximize coherence of the desired signal but to maximize rejection of unwanted mixing images at and near the frequency of the desired signal. This is illustrated for phase error correction in FIG. **16**. An undesired Q phasor UQP is already displaced by amount "a" toward an exact opposite phase relation with the I phasors IP. The Hilbert transform is accordingly employed to rotate the undesired Q phasor UQP and the desired Q phasor DQP by a phase correction amount PC such that PC=90°−a. This rotation moves UQP into direct phase opposition to the I phasors IP, while DQP is not completely phase corrected, being out of phase by amount "a" plus amount "b."

All error correction between the I and Q channels is preferably implemented by running a characterization of each device upon completion of device fabrication, and then storing desired correction factors in a memory associated with the digital portion of the device. Other techniques such as techniques to continuously detect and correct such errors may also be employed, if desired. Temperature sensing capability may also be provided if desired, such that correction factors may be dependent on temperature for optimized image rejection under various climatic conditions.

Allpass filter **40** is designed with nominal group delay equal to the group delay of Hilbert transform **38**. The Hilbert transform **38** or the allpass filter **40** is also enabled to change the sign of its output, in order to switch from rejecting the image of the upper high frequency spectrum of interest to rejecting the image of the lower high frequency spectrum of interest, and vice versa. As explained above, this switching, combined with the correct step size S of the local oscillator **24** and with an appropriately sized and located digital tuning range and/or near-baseband passband, results in twice the local oscillator step size S that would otherwise be possible for a given tuning range or a given channel spacing of a channelized spectrum.

Particularly for the embodiment of the device **10** in FIG. **8** employing the alternate signal paths I_a and Q_a, group delay correction may also be performed in the Hilbert transform pair, if desired, by providing the allpass filter **40** with variable coefficients corresponding to variably offset samples of the sinc function. (The sinc function is defined as:)

$$\sin c(x) = \begin{cases} x = 0: & 1 \\ x \neq 0: & \dfrac{\sin(x)}{x} \end{cases}$$

For zero time shift, the coefficients are given by samples of the sinc function at zero+nπ with n an integer, which results in values of zero everywhere except at n=0, where the sinc function returns a value of 1. Coefficients for a time offset equal to ¼ of one sample at any given sampling frequency may be generated by sampling the sinc function at nπ/4, the central 7 samples of which are shown in FIG. **17**. These seven samples give seven coefficient values, with nπ/4 more obtainable as desired by extending the sinc function sampling further in both directions. An appropriate window (such as Hamming, Blackman, or Kaiser) should, of course, be applied to the coefficient values.

Table III below contains Matlab® code for generating the coefficients for the modified Hilbert transform **38** of FIG. **8**. In the code, fs is the variable for the sampling frequency entering the Hilbert transform and is set to 128 kHz. The fbw variable represents the bandwidth of the signal of interest, in this case set to 6400 for an 8000 bps QPSK square-root raised-cosine digital signal with an excess bandwidth setting (Beta) of 0.6. The fref variable is a reference frequency of the local oscillator equal to the local oscillator step size which is in this case 42.67 kHz. The fref variable is used in this code to define the passband of the transform according the preferred embodiment. Nh is the filter order, set to 16, resulting in 17 filter taps. (Note that this code applies the transform to the Q channel rather than the I channel—either is fine, as long as the other channel has a 0° allpass.)

After the Hilbert transform pair **38**, **40** of the device **10** of FIG. **8**, the signals from the I and Q channels are combined by adder **64** resulting in a real, image-rejected near-baseband signal. This signal is fed to a variable band-pass decimating filter **66**.

The variable band-pass decimating filter **66** is designed by first designing a prototype filter to have a passband of width W straddling DC, where W is the bandwidth of the desired

signal. Similarly to the preferred embodiment of decimating filters **34** and **36**, stopbands also of width W are defined for the prototype filter only at locations aliased to the passband by the decimation sampling rate R_D. Transition bands may again be left unconstrained during filter design. To provide adequate transition band width while preventing undesired aliasing, R_D must be somewhat greater than W.

Once the prototype filter coefficients are obtained, the position of the passband and the stop bands are varied as desired by multiplication of the filter coefficients by a complex exponent to select from the near-baseband passband a desired signal spectrum of width W. The aliasing caused by the decimation then translates the selected spectrum to within $R_D/2$ of baseband. The variable band-pass decimating filter thus performs a tuning function with a resolution of $R_D/2$.

Matlab® code for determining the coefficients of a filter useable as the variable bandpass decimating filter **66** is presented below in tables IV and V. The code in table IV determines the coefficients for a prototype filter with a passband at DC of width W and seven stopbands of width W at intervals of R_D to either side of DC. In the code in table IV, N is the filter order and is set to 63, resulting in 64 FIR filter taps, providing adequate interference attenuation with a power of 2 length which is typically more easily implemented in a DSP. R is the relative emphasis on stopband performance relative to passband performance and is set to 50, resulting in good stopband rejection with reasonable levels of passband ripple. Variable **f1** represents the sampling frequency at the Hilbert transform pair and is set to 128 kHz. Variable **f2** represents the output sampling frequency of filter **66** and is set to 16 kHz, giving a decimation ratio of M=8. Variable fbw represents the bandwidth of the signal of interest, in this case a 8000 bps QPSK square-root raised-cosine digital signal with an excess bandwidth setting (Beta) of 0.6.

The code in table V adjusts the coefficients generated in the code in table IV, which are contained in variable b. Variable fs is the sampling frequency at the Hilbert transform pair, set to 128 kHz here. Variable fshift is the frequency to which the passband of the filter **66** is to be shifted.

Final fine-tuning is performed after filter **66** by fine-shifting by mixing with a complex exponential signal. This fine shifting brings the desired signal to baseband from the location within $R_D/2$ of baseband to which it was aliased by filter **66**. The complex exponential signal is supplied by a digital quadrature local oscillator **68** and mixed with the complex signals by digital mixer **70**. The complex signals are then filtered and decimated at a decimation rate of M=2 by filters **72**, **74** matched to the pulse of the desired signal, which filters reject frequencies in the transition bands of the variable bandpass decimating filter and in regions aliasing into the desired signal. Significant suppression of transition bands is thus effectively postponed until the signal reaches filters **72**, **74**, at which point the sampling frequency has been reduced to R_D=R/8, and the spectrum of interest has been reduced to within $R_D/2$ of baseband, allowing for relatively efficient sharp-cutoff filtering. The resulting signals are then demodulated by a quadrature demodulator. For other types of signals, other typical demodulation procedures and devices may be used after the fine-shifting operation.

Simulated frequency response curves for the embodiment of the device **10** of FIG. **8** not employing oversampling are seen in FIGS. **18–20**.

FIG. **18** shows the continuous-time filtering frequency response curve CT together with the Hilbert transform frequency response curve HT. The CT curve shows the desired attenuation of frequencies near DC, together with anti-aliasing low pass filtering with a transition band from about 2.25 to about 3. Beginning the transition band above 2, the upper boundary of the near-baseband passband in this embodiment, avoids including the corner of the lowpass filter within the near-baseband passband, thereby avoiding significant group delay variations which can increase intersymbol interference. The beginning of the stopband at 3 could actually be relaxed, allowing the stopband to begin as late as 4 (minus F_A) without resulting in aliasing into the near-baseband passband at 1 to 2, given a sampling frequency of R=6 as preferred with a 1 to 2 near-baseband passband. The HT curve shows rejection of the mixing image at −1 to −2 on the x axis.

FIG. **19** shows a variable passband decimating filter frequency response curve VDF and a matched filter frequency response curve MF. The VDF curve shows a passband centered at 1 with seven stopbands to either side at intervals of 0.75=R/8=R_D. The matched filter frequency response curve MF is shown aligned with the VDF curve to reject signals within the transition band of the VDF curve. Translation to baseband is not shown.

FIG. **20** shows the system frequency response resulting from the cascade of the frequency responses shown in FIGS. **18** and **19**.

FIG. **21** shows an envelope detector output resulting from simulated envelope detection of the simulated output of the device **10** of FIG. **8**. Intersymbol interference is −28.55 dB, well below the maximum allowable in most digital modulation schemes.

FIGS. **22–25** show simulated frequency response curves for the embodiment of the device **10** of FIG. **8** employing oversampling, but with M=16 (16 times oversampling) for illustration purposes rather than M=32 as currently most preferred.

FIG. **22** shows the continuous-time filtering response curve CT, with the very relaxed upper transition bands employable with oversampling. The frequency curve HT response of the Hilbert transform and the frequency response curve DF of the decimating filters **34**, **36** are also shown. The DF curve shows the desired passband at about 1 to 2 on the x axis with stopbands of the same width repeating at intervals of 3 on either side.

FIG. **23** shows the curves of FIG. **22** on a smaller scale. The unconstrained transition bands of the DF curve may be seen, as well as the image rejection of the HT curve at 1 to 2.

FIGS. **24** and **25** show a simulated system frequency response resulting from the embodiment of device **10** in Figure using oversampling at M=16. Attenuation of unwanted signals and noise meets the requirements of most commercially available mobile radio equipment (about 75 dB).

While the above described preferred embodiments are only the presently most preferred embodiments, certain additional general advantages of the present invention may be seen therein.

The particular division of functions between analog and digital portions of the device **10** allows sufficiently relaxed requirements for both the analog and digital portions that each can be implemented individually on a single integrated circuit. The processing power and memory requirements of the embodiment not employing oversampling are low enough to allow implementation of the entire digital portion in a current general purpose DSP. Even the oversampling embodiment may potentially be implemented in next-

generation general purpose DSPs, or in an ASIC with no more complexity or power requirements than a general purpose DSP, or in a current DSP with a discrete component or two for performing the delta-sigma modulation. Single chip implementation of the entire device is even possible with adequate shielding of the analog portions from digital portion noise.

While not required, all of the filters in the digital domain may be implemented as single stage filters due in part to the location of the near-baseband passband of the present invention, and to the use of non-constrained transition bands. Single stage filtering is advantageous in the variable bandpass decimating filter **66**, since only a single stage of coefficients must be varied to alter the passband location. The passband could thus be varied in real time to follow variations in a particular signal of interest. Single stage filtering is also very advantageous in the oversampling decimating filters **34**, **36** because it allows elimination of multipliers and implementation with adders only.

The preferred use of a modified type III Hilbert transform allows particularly easy and efficient correction of phase errors between the quadrature I and Q signals. The use of a one bit wide signal path in the oversampling embodiment also allows easy correction of group delay differences between the I and Q signals.

As known to those of skill in the art, modulation and transmission of RF signals may be performed essentially by reversing the demodulation process. The device of the present invention may accordingly be adapted by those of skill in the art for use as a transmitter/receiver, if desired. In this case, the signal flow shown in FIG. **8** would be reversed. I and Q signals from a baseband modulator would be sent through transmit stages analogous to the receive stages shown. For example, a splitter would be substituted for the adder, interpolating filters substituted for the decimating filters, and digital to analog converters substituted for the analog to digital converters. The Hilbert transform pair and local oscillator(s) would function for transmission in the same manner as they function for reception. This substitution of analogous transmit stages for receive stages should be readily apparent to those of skill in the art.

Having illustrated and described the principles of the invention in a preferred embodiment, it should be apparent to those skilled in the art that the embodiment can be modified in arrangement and detail without departing from such principles. For example, while device **10** of a preferred embodiment discussed above includes a digital portion that corrects phase and amplitude errors, an RF receiver according to various aspects of the invention can include just the following: (1) a local oscillator coarse-tunable in steps of step size S for producing a local oscillator signal; (2) continuous-time quadrature mixers for quadrature mixing with the local oscillator signal an incoming channel of interest from a channelized spectrum having channel spacing C, to produce I and Q signals approximately in quadrature, thereby translating the channel of interest to a near-baseband passband; and (3) a digital signal processing device for rejecting an unwanted mixing image within the near-baseband passband. The digital signal processing device can translate any one of N channels of the channelized spectrum from the near-baseband passband to baseband, where N is a constant positive integer equal to the number of channels of the channelized spectrum contained simultaneously within the near-baseband passband, wherein S is at least 2(N−1)C. Alternatively, it can translate to baseband a signal of interest within the near-baseband passband, said signal of interest within the near baseband passband having essentially any center frequency within a range extending from a lower frequency F1 to an upper frequency F2 inclusive, wherein S is about 2(F2−F1). In view of the many possible embodiments to which the principles of the invention may be applied, it should be recognized that the illustrated embodiment is only a preferred example of the invention and should not be taken as a limitation on the scope of the invention. Rather, the invention is defined by the following claims. I therefore claim as my invention all such embodiments that come within the scope and spirit of the following claims.

TABLE I

```
%       Define the delta-sigma oversampling frequency
%       from the oversampling ratio
%       and the discrete-time system sampling frequency
%       (input to the Hilbert transform pair)
        fds = fs * Mds;
%       Define weight of passband as 1/R and weight of all
%       stopbands as 1
        wt = [1/R ones(1,floor(Mds)-1];
%       Define passband of filter
        f = [fs/6 fs/3 ] ;
        m = [1 1 ] ;
%       Define stopbands of filter only at frequency regions
%       which would alias into passband
for k = 1:ceil(Mds)-1,
        %   Stopbands fs/2 apart because real filter
        f(2*k+1:2*k+2) = [ -fs/12 fs/12] + . . .
        (fs/2*k+fs/4)*ones(1,2) ;
        m(2*k+1:2*k+2) = [ 0 0] ;
end
%       Adjust for 1 = Nyquist freg.
        f = f ./ (0.5*fds) ;
%       Compute filter using Remez exchange algorithm
        bds = remez (Nds,f,m,wt) ;
```

TABLE II

LOOK-UP TABLE COEFFICIENTS			
54a	54b	54c	54d
164	4454	−6504	7308
222	4538	−7064	9174
276	4602	−7626	11106
348	4646	−8146	13098
426	4668	−8638	15138
514	4664	−9092	17224
608	4636	−9508	19342
710	4582	−9876	21486
820	4498	−10194	23648
936	4384	−10456	25816
1060	4238	−10656	27980
1190	4060	−10790	30134
1328	3850	−10852	32264
1472	3604	−10838	34362
1622	3324	−10742	36418
1778	3008	−10562	38420
1940	2658	−10292	40360
2106	2272	−9930	42228
2276	1852	−9474	44016
2450	1398	−8920	45712
2624	912	−8266	47310
2802	394	−7510	48798
2978	−156	−6652	50172
3154	−732	−5692	51424
3328	−1334	−4630	52546
3498	−1960	−3468	53532
3664	−2504	−2204	54378
3822	−3266	−844	55076
3972	−3940	612	55630
4112	−4624	2160	56028
4240	5304	3794	56288
4354	−5996	5512	56372

TABLE III

```
%       Define the zero degree delay as a unit sample in
%       the center of an FIR filter of equal length to the
%       90 degree Hilbert transform filter. Only the
%       samples before the unit sample need to be
%       implemented in practice.
        bhI = [zeros (1, (Nh−1)/2) 1 zeros (1, (Nh−1)/2)]
%       Define passband of the Hilbert transform
        fhilb = [(fref/2 − fbw/2) (fref + fbw/2)] / (fs/2) ;
%       Define the 90 degree Hilbert transform using Remez
%       exchange algorithm
        bhQ = remez (Nh,fhilb, [1 1], 'Hilbert') ;
%       Scale amplitude of Q channel coefficients for gain
%       imbalance compensation
        bhQv = QG .* bhQ .* SB;
%       Adjust amplitude of center Q channel coefficient to
%       vary phase from 90 degrees. This provides phase
%       imbalance compensation.
        bhQv ((Nh/2) +1) = tan((2*pi) /360 * QP) ;
```

TABLE IV

```
function b = dbf (N,R,f1,f2,fbw)
%       b = dbf (N,R,f1,f2,fbw)
%       Decimating Bandpass Filter
%       Nth-order bandpass FIR with decimation and
%       asymmetrical frequency response
%       R parameter (50 suggested) determines relative
%       importance of passband and stopbands
%       f1 is input sampling rate
%       f2 is output sampling rate
%       fbw is 2-sided bandwidth of interest
%       Calculate cutoff frequency of prototype LPF (1/2 BW
%       of interest)
        fc = fbw/2;
%       Iterate to get all stopband regions except at f=1
        M = f1/f2;
%       Parks McClellan: define bands
        ----------------------------------------------------
%       Define passband of filter
        f = [ 0 fc ] ;
        m = [ 1 1 ] ;
%       Define weight of stopband vs. passband based on R
        parameter
%       50 gives Rp<0.01dB with Rs<−100dB
%       Define weight of passband as 1/R and weight of all
%       stopbands as 1
        wt = [ 1/R ones (1,floor(M/2))] ;
%       Define stopbands of filter only at frequency regions
%       which would alias into passband
        for k = 1:ceil(M/2) −1,
            f(2*k+1:2*k+2) = [ −fc fc ] + f2*k*ones (1,2);
            m(2*k+1:2*k+2) = [ 0 0 ] ;
end
%       If M is even, append a fixed stopband at Nyquist
%       frequency
if ceil (M/2) = =M/2
            f(M+1 :M+2) = [ f1/2-fc f1/2 ] ;
            m(M+1 :M+2) = [ 0 0 ] ;
end
%       Adjust for 1 = Nyquist freg.
        f = f ./ (05*f1);
%       Compute filter using Remez exchange algorithm
        if N >=3
            b = remez(N,f,m,wt) ;
        else
            %Order less than 3 is defined to be just a
            %zero-order
            %allpass
              b = 1
end
%       Define transfer function denominator coefficients as
%       simply a one followed by zeroes
        a = [1 zeros (1,max(size(b)) −1))] ;
```

TABLE IV-continued

```
%       Make sure H(0) =1
        k = freqz(b,a,linspace(0,pi)) ;
        b = b ./ max(abs(k)) ;
%end of function
end
```

TABLE V

```
n = 1:length(b) ;
b = b . * exp(i*2*pi(fshift/fs)*n)) ;
```

I claim:

1. A method for tuning high frequency signals, the method comprising:

(a) using a local oscillator coarse-tunable in steps of size S, generating a first local oscillator signal at a selected frequency above a lower high frequency spectrum of interest and below an upper high frequency spectrum of interest; and

(b) at the selected local oscillator frequency, switching between:

(1) providing spectrum coverage within the lower high frequency spectrum of interest and not the upper high frequency spectrum of interest; and

(2) providing spectrum coverage within the upper high frequency spectrum of interest and not the lower high frequency spectrum of interest.

2. The method of claim **1** wherein the lower and upper high frequency spectra of interest are separated from the first frequency by a predetermined offset and are separated from each other by twice the predetermined offset.

3. The method of claim **1** further comprising providing a second local oscillator signal having the first frequency and being approximately in quadrature to the first local oscillator signal.

4. The method of claim **1** wherein:

(a) each edge of the lower and upper high frequency spectra of interest nearest the local oscillator signal is separated from the local oscillator signal by a frequency $F1-F_A$, wherein F_A is given frequency adjustment;

(b) each edge of the lower and upper high frequency spectra of interest farthest from the local oscillator signal is separated from the local oscillator signal by a frequency $F2+F_A$; and

(c) $S=2(F2-F1)$.

5. The method of claim **4** wherein $F1=k(F2-F1)$ where k is a positive integer

6. The method of claim **5** further comprising switching the local oscillator signal from the first frequency to a second frequency, wherein the second frequency is one local oscillator step S away from the first frequency.

7. A method for tuning a signal from a channelized spectrum having a predetermined channel spacing, the method comprising

(a) mixing a signal of interest having a predetermined maximum bandwidth with a first local oscillator signal; wherein

(b) the first local oscillator signal has a frequency that (1) is an integer multiple of the channel spacing and (2) is selected to frequency translate the signal of interest to within a near-baseband passband whose lower edge is spaced from DC by at least about the maximum bandwidth of the signal of interest;

whereby problems associated with 1/f noise, DC offsets, and self-mixing products are avoided or substantially diminished.

8. The method of claim **7** further comprising mixing the signal of interest with a second local oscillator signal having the first frequency and being approximately in quadrature with the first local oscillator signal.

9. The method of claim **7** further comprising coarse-tuning the local oscillator signal by one local oscillator step from the first frequency to a second frequency an integral number of channel spacings from the first frequency.

10. The method of claim **9** wherein the second frequency is two channel spacings from the first frequency.

11. The method of claim **7** wherein:

(a) the signal of interest lies within one of an upper high frequency spectrum of interest and a lower high frequency spectrum of interest; and

(b) the method further comprises providing spectrum coverage within one of the high frequency spectra of interest and not the other.

12. The method of claim **11** further comprising switching between:

(a) providing spectrum coverage within the lower high frequency spectrum of interest and not the upper high frequency spectrum of interest; and

(b) providing spectrum coverage within the upper high frequency spectrum of interest and not the lower high frequency spectrum of interest.

13. The method of claim **7** wherein the spacing of the lower edge of the near-baseband passband from DC is greater than the passband's width.

14. The method of claim **13** wherein the spacing of the lower edge of the near-baseband passband from DC is about twice the passband's width.

15. Apparatus for tuning, from a channelized spectrum having a predetermined channel spacing, a signal of interest having a predetermined maximum bandwidth, the apparatus comprising:

(a) a local oscillator configured to generate a local oscillator signal at a frequency that is an integer multiple of the channel spacing; and

(b) a mixer responsive to the local oscillator signal and the signal of interest, wherein the mixer frequency translates the signal of interest; wherein

(c) the frequency-translated signal of interest falls within a near-baseband passband spaced from DC by a frequency offset of at least about the maximum bandwidth of the signal of interest;

whereby problems associated with 1/f noise, DC offsets, and self-mixing products are avoided or substantially diminished.

16. The apparatus of claim **15** wherein the spacing of the lower edge of the near-baseband passband from DC is greater than the passband's width.

17. The apparatus of claim **16** wherein the spacing of the lower edge of the near-baseband passband from DC is about twice the passband's width.

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